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VOLUME 17 OCTOBER, 1929 NUMBER 10

PROCEEDINGS
of
The Institute of Radio Engineers

7267

General Information and Subscription Rates on Page 1666

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Institute of Radio Engineers

Forthcoming Meetings

CINCINNATI SECTION
Cincinnati, Ohio, October 17, 1929

CLEVELAND SECTION
Cleveland, Ohio, October 18, 1929

CONNECTICUT VALLEY SECTION
Hartford, Conn., October 14, 1929

DETROIT SECTION
Detroit, Mich., October 16, 1929

NEW YORK MEETING
New York, N. Y., November 6, 1929

PITTSBURGH SECTION
Pittsburgh, Penna., October 15, 1929

TORONTO SECTION
Toronto, Canada, October 15, 1929

WASHINGTON SECTION
Washington, D. C., October 10, 1929

Eastern Great Lakes District Convention

November 18-19, 1929, Rochester, N. Y.

Sponsored by the

Rochester, Buffalo-Niagara, Cleveland
and Toronto Sections

All members of the Institute are invited to participate in the very interesting two-day program being arranged

(SEE PAGE 1671 OF THIS ISSUE FOR COMPLETE PROGRAM)

PROCEEDINGS OF

The Institute of Radio Engineers

Volume 17 October, 1929 Number 10

Board of Editors, 1929
WALTER G. CADY, *Chairman*
STUART BALLANTINE G. W. PICKARD
RALPH BATCHER L. E. WHITTEMORE
CARL DREHER W. WILSON

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On the Relation between Long-Wave Reception and Certain Terrestrial and Solar Phenomena, by K. SREENIVASAN 1793
The Significance of Observations of the Phase of Radio Echoes, by G. BREIT 1815
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Single-Wire Transmission Lines for Short-Wave Antennas, by W. L. EVERITT AND J. F. BYRNE 1840
Circuit Tuning by Wave Resonance and Applications to Radio Reception, by LOUIS COHEN 1868
Wave Resonance Tuning and Application to Radio Transmission, by WILLIAM R. BLAIR AND LOUIS COHEN 1893
Discussion on "Radio Direction-Finding by Transmission and Reception," by R. L. SMITH-ROSE 1897
Discussion on "Microphonic Improvement in Vacuum Tubes," by ALAN C. ROCKWOOD AND WARREN R. FERRIS 1899
Book Review, "Radio Encyclopedia" 1903
Books Received 1903
Monthly List of References to Current Radio Literature 1905
Contributors to This Issue 1911

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The Institute of Radio Engineers

GENERAL INFORMATION

The PROCEEDINGS of the Institute is published monthly and contains papers and discussions thereon submitted for publication or for presentation before meetings of the Institute or its Sections. Payment of the annual dues by a member entitles him to one copy of each number of the PROCEEDINGS issued during the period of his membership.

Subscription rates to the PROCEEDINGS for the current year are received from non-members at the rate of $1.00 per copy or $10.00 per year. To foreign countries the rates are $1.10 per copy or $11.00 per year.

Back issues are available in unbound form for the years 1918, 1920, 1921, 1922, and 1926 at $9.00 per volume (six issues) or $1.50 per single issue. Single copies for the year 1928 are available at $1.00 per issue. For the years 1913, 1914, 1915, 1916, 1917, 1918, 1924, and 1925 miscellaneous copies (incomplete unbound volumes) can be purchased for $1.50 each; for 1927 at $1.00 each. The Secretary of the Institute should be addressed for a list of these.

Discount of twenty-five per cent on all unbound volumes or copies is allowed to members of the Institute, libraries, booksellers, and subscription agencies.

Bound volumes are available as follows: for the years 1918, 1920, 1921, 1922, 1925, and 1926 to members of the Institute, libraries, booksellers, and subscription agencies at $8.75 per volume in blue buckram binding and $10.25 in morocco leather binding; to all others the prices are $11.00 and $12.50, respectively. For the year 1928 the bound volume prices are: to members of the Institute, libraries, booksellers, and subscription agencies, $9.50 in blue buckram binding and $11.00 in morocco leather binding; to all others, $12.00 and $13.50, respectively. Foreign postage on all bound volumes is one dollar, and on single copies is ten cents.

Year Books for 1926, 1927, and 1928, containing general information, the Constitution and By-Laws, catalog of membership etc., are priced at seventy-five cents per copy per year.

Contributors to the PROCEEDINGS are referred to the following page for suggestions as to approved methods of preparing manuscripts for publication in the PROCEEDINGS.

Advertising rates for the PROCEEDINGS will be supplied by the Institute's Advertising Department, Room 802, 33 West 39th Street, New York, N. Y.

Changes of address to affect a particular issue must be received at the Institute office not later than the 15th of the month preceding date of issue. That is, a change in mailing address to be effective with the October issue of the PROCEEDINGS must be received by not later than September 15th. Members of the Institute are requested to advise the Secretary of any change in their business connection or title irrespective of change in their mailing address, for the purpose of keeping the Year Book membership catalog up to date.

The right to reprint limited portions or abstracts of the papers, discussions, or editorial notes in the PROCEEDINGS is granted on the express condition that specific reference shall be made to the source of such material. Diagrams and photographs published in the PROCEEDINGS may not be reproduced without making special arrangements with the Institute through the Secretary.

It is understood that the statements and opinions given in the PROCEEDINGS are views of the individual members to whom they are credited, and are not binding on the membership of the Institute as a whole.

All correspondence should be addressed to the Institute of Radio Engineers, 33 West 39th Street, New York, N. Y., U. S. A.

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33 West 39th St., New York, N. Y.

SUGGESTIONS FOR CONTRIBUTORS TO THE PROCEEDINGS

Preparation of Paper

Form—Manuscripts may be submitted by member and non-member contributors from any country. To be acceptable for publication manuscripts should be in English, in final form for publication, and accompanied by a summary of from 100 to 300 words. Papers should be typed double space with consecutive numbering of pages. Footnote references should be consecutively numbered and should appear at the foot of their respective pages. Each reference should contain author's name, title of article, name of journal, volume, page, month, and year. Generally, the sequence of presentation should be as follows: statement of problem; review of the subject in which the scope, object, and conclusions of previous investigations in the same field are covered; main body describing the apparatus, experiments, theoretical work, and results used in reaching the conclusions and their relation to present theory and practice; bibliography. The above pertains to the usual type of paper. To whatever type a contribution may belong, a close conformity to the spirit of these suggestions is recommended.

Illustrations—Use only jet black ink on white paper or tracing cloth. Cross-section paper used for graphs should not have more than four lines per inch. If finer ruled paper is used, the major division lines should be drawn in with black ink, omitting the finer divisions. In the latter case, only blue-lined paper can be accepted. Photographs must be very distinct, and must be printed on glossy white paper. Blueprinted illustrations of any kind cannot be used. All lettering should be $^3/_{16}$ in. high for an 8 x 10 in. figure. Legends for figures should be tabulated on a separate sheet, not lettered on the illustrations.

Mathematics—Fractions should be indicated by a slanting line. Use standard symbols. Decimals not preceded by whole numbers should be preceded by zero, as 0.016. Equations may be written in ink with subscript numbers, radicals, etc., in the desired proportion.

Abbreviations—Write a.c. and d.c., (a-c and d-c as adjectives), kc, μf, μμf, e.m.f., mh, μh, henries abscissas, antennas. Refer to figures as Fig. 1, Figs. 3 and 4, and to equations as (5). Number equations on the right in parentheses.

Summary—The summary should contain a statement of major conclusions reached, since summaries in many cases constitute the only source of information used in compiling scientific reference indexes. Abstracts printed in other journals, especially foreign, in most cases consist of summaries from published papers. The summary should explain as adequately as possible the major conclusions to a non-specialist in the subject. The summary should contain from 100 to 300 words, depending on the length of the paper.

Publication of Paper

Disposition—All manuscripts should be addressed to the Institute of Radio Engineers, 33 West 39th Street, New York City. They will be examined by the Committee on Meetings and Papers and by the Editor. Authors are advised as promptly as possible of the action taken, usually within one month.

Proofs—Galley proof is sent to the author. Only necessary corrections in typography should be made. *No new material is to be added.* Corrected proofs should be returned *promptly* to the Institute of Radio Engineers, 33 West 39th Street, New York City.

Reprints—With the galley proof a reprint order form is sent to the author. Orders for reprints must be forwarded promptly as type is not held after publication.

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J. H. DELLINGER
Member of Board of Direction, 1929

J. H. Dellinger was born in Cleveland, Ohio, July 3, 1886. He was educated at Western Reserve (1903–07), George Washington University (A.B. degree, 1908), and Princeton University (1912–13, Ph.D. degree in 1913). He was an instructor in the physics department of Western Reserve in 1906–07.

Dr. Dellinger joined the staff of the Bureau of Standards as physicist in 1905. From 1919 to date he has been Chief of the Bureau's Radio Section. He has been particularly active in special government service. In 1912 he was a U. S. delegate to the Interallied Technical Conference on Radio Communication at Paris and member of technical staff of the Conference on Limitation of Armament and Far East Problems, at Washington. In 1922–23 he served as secretary of the U. S. Government Inter-department Radio Advisory Committee. He was a member of all four National Radio Conferences (1922–25), and since 1924 has been chairman of the Committee on Radio Apparatus, Federal Specifications Board. Dr. Dellinger was on the technical advisory staff of the International Radio Conference at Washington in 1927 and has recently left for The Hague, where he is one of the technical advisors to the American Delegation of the International Technical Consulting Committee.

Dr. Dellinger was vice-president of the Institute in 1924 and president in 1925. In 1928 he was Chairman of the Committee on Meetings and Papers and is the present chairman of the Committee on Standardization. He was elected a member of the Board of Direction of the Institute by the membership in 1928 to serve until 1931. He is a Fellow in the Institute of Radio Engineers and the American Physical Society.

INSTITUTE NEWS AND RADIO NOTES

Eastern Great Lakes District Convention

An innovation in general meetings of Institute members is to be provided at the forthcoming Eastern Great Lakes District Convention which is to be held in Rochester, New York, on November 18–19, 1929.

The Convention, to which the entire membership of the Institute is cordially invited, is sponsored by the Rochester, Buffalo-Niagara, Toronto, and Cleveland sections of the Institute. A two-day program of diversified and interesting activities including four technical papers of timely importance and three inspection trips has been arranged.

The headquarters will be at the Sagamore Hotel, where all technical sessions and the banquet on the second evening will take place.

Registration cards will be mailed from the Institute office in advance of the meeting. Members contemplating attending the Convention should return these cards promptly to the Convention Headquarters.

Program

Sunday, November 17th—3:00 P.M. to 6:00 P.M.—Registration at Convention Headquarters, Sagamore Hotel.

Monday, November 18th—8:00 P.M.—Registration at Sagamore Hotel, Convention Headquarters.

10: A.M. to 12:30 P.M.—Opening technical session. Address of welcome by E. C. Karker, chairman of the Rochester section. Technical papers: "Considerations in Screen-Grid Receiver Design," by W. A. McDonald, Hazeltine Corporation, Hoboken, N. J., and "What Executives Expect of Engineers," by I. G. Maloff, Valley Appliance Corp., Rochester.

2:00 P.M.—Inspection trip to Kodak Park and Valley Appliance Corporation.

8:00 P.M.—Technical session: "Ultra High-Frequency Transmission and Reception," by A. Hoyt Taylor, Naval Research Laboratory, Washington, D. C.

Tuesday, November 19th—9:30 to 12:30 P.M. Technical session: "A Broadcast Receiver for Special Purposes," by Paul O. Farnham, Radio Frequency Laboratories, Boonton, N.J. and "Standardization in the Radio Vacuum-Tube Field," by W. C. White, Research Laboratory, General Electric Company, Schenectady.

2:00 P.M.—Inspection trip to Stromberg-Carlson Plant.

6:30 P.M.—Banquet at Sagamore Hotel. Speaker, H. B. Richmond, President of the Radio Manufacturers' Association.

The following committees have been appointed and are arranging the entire Convention:

Executive: Virgil M. Graham, chairman; publicity, R. A. Hackbusch, chairman; transportation-accommodations, J. Eichman, chairman; fellowship, J. A. Victoreen, chairman; trips, A. L. Schoen, chairman; entertainment, E. C. Karker, chairman; registration, A. E. Soderholm, chairman; exhibits, I. G. Maloff, chairman.

September Meeting of Board of Direction

At the meeting of the Board of Direction of the Institute held at 2:00 P.M. on September 4th, the following board members were present: J. H. Dellinger (acting chairman), Melville Eastham, treasurer; John M. Clayton, secretary; Arthur Batcheller, Ralph Bown, Alfred N. Goldsmith, R. H. Marriott, and L. E. Whittemore.

The following were transferred or elected to the higher grades of membership in the Institute: transferred to the grade of Fellow: C. W. Latimer; transferred to the grade of Member: V. D. Landon, C. B. Mirick, E. R. Pfaff, Paul M. Segal, and Clinton H. Suydam; elected to the grade of Member: J. H. Barron, J. A. Beveridge, R. J. Bird, Louis G. Caldwell, A. L. Duval d'Adrian, Lionel Harvey, C. M. Johnson, R. H. Lyne, George McElrath, R. A. McLennan, L. Osiatinsky, H. A. Pidgeon, E. G. Ramirez, D. E. Replogle, Hans Roder, W. G. Russell, B. A. Schwarz, A. C. Shaw, K. H. Thow, Glen E. West, R. M. Wilmotte, W. L. Winner, E. C. Woodruff.

One hundred and eighteen Associate members and fourteen Junior members were elected.

The Board approved the tentative program of the 1930 Convention as submitted by the Convention Committee, deciding upon August 18–21 as the most suitable dates.

The Board approved the petition for the formation of a Cincinnati section of the Institute. W. W. Boes and R. H. Langley were prominent in the organization work which led to the petition. The petition bore the signatures of the following members of the Institute: C. D. Barbulesco, Carl L. Flory, C. F. Fraasa, C. C. Dixon, R. M. Blair, R. P. Glover, F. J. Franzwa, W. S. Hannon, C. W. Peterson, H. S. Meyrose, E. J. Bussard, H. E. Youngs, C. E. Kilgow, D. D. Israel, J. A. Chambers, F. Clifford Estey, E. T. Flewelling, H. J. Loftis, W. K. Aughenbaugh, R. H. Langley, W. W. Boes, L. A. Kellogg, L. A. Hendricks, John Gerstle, and R. S. Copp.

Nomination of 1930 Officers and Managers

At the September 4th meeting of the Board of Direction the following were placed in nomination for 1930 elective officers and Board members:

For President: R. A. Heising, engineer, technical staff, Bell Telephone Laboratories, New York, N. Y. Member of the Board of Direction, 1927–29; Chairman, Committee on Admissions, 1929. Fellow of the Institute.

For Vice President: A. G. Lee, staff engineer in charge of radio section, engineering department, General Post Office, London, England. Member, Committee on Admissions, 1928–29. Fellow of the Institute.

For Managers (three-year terms)

J. V. L. Hogan, consulting engineer, New York, N. Y. Vice-president of the Institute, 1916–19, president, 1920. Member, Board of Direction, 1913–15, 1921–29. Member, Committee on Broadcasting, 1928–29. Fellow of the Institute.

R. H. Marriott, consulting engineer, New York, N. Y. President of the Institute, 1912. Member, Board of Direction, 1914–16, 1920–22, 1926–29. Member, Committee on Broadcasting, 1928, vice-chairman, 1929; Chairman, Committee on Constitution and Laws, 1928–29. Fellow of the Institute.

Ballots will be in the mails on or before October 20, 1929.

Institute Representative at International Conference

The Institute has been invited by the Department of State to send a representative to the first meeting of the International Technical Consulting Committee on Radio Communication held at The Hague in September, 1929. The International Technical Consulting Committee is charged with the study of technical and allied problems which relate to international radio communication. Its functions are to give advice on questions submitted by the various governments. The Dutch government organized the first meeting of the Committee and prepared the program of work.

The Institute has appointed Dr. W. Wilson as its representative at the Conference. Dr. Wilson sailed for The Hague early in September.

The American delegation is composed of the following members: Major-General Charles McK. Saltzman, Major-General George S. Gibbs, and Captain S. C. Hooper. Among the technical assistants appointed for the delegation are J. H. Dellinger, C. B. Jolliffe, Lieut. Comdr. T. A. M. Craven, G. C. Gross, and K. B. Warner. The representatives of American communication companies attending the conference include Lloyd Espenschied, W. Wilson, J. Nathansonn, L. A. Briggs, Colonel S. E. Reber, A. Y. Tuel, H. H. Buttner, W. E. Beakes, R. M. Heintz, C. J. Pannill, J. W. Swanson, L. G. Caldwell, and H. C. Leuteritz.

Announcement of Civil Service Examination

The U. S. Civil Service Commission announces an open competitive examination for the position of assistant inspector (radio enforcement) at a salary of $2,400 a year. Competitors will be rated on the following subjects: theoretical and practical questions on radio and electrical engineering, 60 per cent; education, training and experience, 40 per cent.

The duties of the assistant radio inspector will be primarily to assist the radio inspector in enforcement of the radio act, and will include the inspection of radio equipment on vessels and at land stations. Members of the Institute interested in this examination should secure Form 2600 from the U. S. Civil Service Commission, Washington, D. C., mentioning examination number 205, and the exact title of examination desired.

Convention Reprints Available

By addressing a letter to the Secretary of the Institute, members will be supplied with one copy of any of the following 1929 Convention reprints, gratis. Mention should be made of the reprint number when writing.

No. 75—Symposium on Technical Problems of Radio Regulation
No. 76—Symposium on Photoradio and Television
No. 77—Nineteen papers presented at the joint meeting of the Institute and the American Section of the International Scientific Radio Union.

Past Issues of the Proceedings

On page XXI of the advertising section of this issue will be found a statement of the back issues of the PROCEEDINGS, both in bound volumes and single copies, on hand for sale at the Institute office.

Institute members contemplating completing their files of back issues are informed that the stock of certain numbers and years is very small, and are advised to place an order with the Institute for those copies required in the near future before the present stock is exhausted.

Members desiring photostat copies of issues of the PROCEEDINGS or individual papers no longer available for sale should communicate with the United Engineering Societies Library, 33 West 39th Street. The price for photostat copies is twenty five cents per double page.

Institute Meetings

NEW YORK MEETING

On Wednesday, September 4th, the first meeting of the fall season was held in the Engineering Societies Building, 33 West 39th Street, presided over by R. H. Marriott. The following papers were presented: "Plate Voltage Supply for Naval Vacuum Tube Transmitters," by Commander E. C. Raguet, Bureau of Engineering, Navy Department, and "'Hot-Cathode Mercury-Vapor Rectifier Tubes," by H. C. Steiner and H. T. Maser of the ResearchLaboratory of the General Electric Company at Schenectady. The first paper was delivered by Lieutenant Commander R. C. Starkey, of the Brooklyn Navy Yard, New York, and the second paper by H. C. Steiner.

Following the presentation of these papers the following participated in their discussion: H. C. Steiner, H. T. Maser, H. E. Hallborg, W. C. White, Robert M. Arnold, Austin Bailey, and Frank R. Stansel.

ATLANTA SECTION

The first meeting of the fall season of the Atlanta section was held on September 4th in the Hotel Winecoff. An informal banquet preceded the technical meeting.

Henry L. Reid presented a paper on "Modern Methods of Servicing Radio Receivers." Phil Bangs presented an informal talk on benefits of I. R. E. membership. Lieutenant S. A. Jones, U. S. N., gave a talk on "Experiences in China."

Mr. Reid's paper described the service problem from 1920 through to 1929 and demonstrated the latest model of a set analyzer, explaining the function of the various circuits. Lieutenant Jones' paper dealt with the trade conditions, customs and possibilities of the Far East as a market for American goods.

Mr. Bangs emphasized the fact that members of the sections of the Institute are interested in all phases of the radio art and that section meetings should offer an opportunity for members to discuss problems arising in their particular field.

PITTSBURGH SECTION

A joint meeting of the Pittsburgh sections of the Institute and the A. I. E. E. will be held on November 12th. Mr. Taylor, of the General Electric Company, will present a paper, "Making Sound Visible and Light Audible."

Committee Work

COMMITTEE ON ADMISSIONS

A meeting of the Committee on Admissions was held at 6:30 P.M. on August 14th in the Western Universities Club, Hotel Biltmore, New York City. The following were present: R. A. Heising, chairman; E. R. Shute, H. S. Smith, A. F. Van Dyck, R. S. Kruse, H. F. Dart, George Lewis, and F. R. Brick.

The committee took action on thirty-three applications for transfer or election to higher grades of membership in the Institute, approving twenty-five of that number.

COMMITTEE ON BROADCASTING

Raymond Guy, of the Research Department of the Radio Corporation of America, New York, N. Y., has been appointed to membership on the Committee on Broadcasting under the chairmanship of L. M. Hull.

COMMITTEE ON MEETINGS AND PAPERS

As foreign members of the Committee on Meetings and Papers to officially represent the Institute in that capacity in their individual countries, President Taylor has appointed the following:

Dr. Alexander Meissner, Germany
Professor E. V. Appleton, England
Dr. Balth. van der Pol, Holland
Prof. V. T. Bashenoff, U. S. S. R
Prof. H. Yagi, Japan

COMMITTEE ON SECTIONS

A meeting of the Committee on Sections was held at 6:30 P.M. on August 29th in the Western Universities Club, New York, N. Y., with attendance as follows: E. R. Shute, chairman; Austin Bailey and D. H. Gage.

The Committee approved the petition from members in the Cincinnati, Ohio, territory, recommending to the Board of Direction that a Cincinnati section be organized.

The Committee completed its revision of the Manual of Section Organization and Operation, and took action on a number of the detailed section operation matters.

SECTIONAL COMMITTEE ON RADIO

Reorganization of the Sectional Committee on Radio of the American Standards Association, of which the I.R.E. and the A.I.E.E. are joint sponsors, is in progress.

The Board of Direction of the Institute has designated L. E. Whittemore, L. M. Hull, and J. V. L. Hogan representatives of the Institute of Radio Engineers on the Sectional Committee on Radio.

Personal Mention

Alfred Crossley has resigned from the Steinite Radio Company to become chief engineer of the Howard Radio Company of Chicago.

M. C. Batsel, until recently in the radio receiver section of the Westinghouse Electric and Manufacturing Company at East Pittsburgh, is now chief engineer of RCA Photophone, Inc., in New York City.

D. J. H. Leitch has joined Radio Industries of Canada, Ltd., as branch manager, with headquarters at Toronto.

C. G. Lemon, for some time in the research and experimental department of Messrs. Lissen, Ltd. of Isleworth, England, is associated with Tungsram Electric Lamp Works of London as technical adviser.

Frank Leslie, formerly assistant chief engineer of Farrand Manufacturing Company, is now recording engineer at Paramount Famous Lasky Corporation in Astoria, N. Y.

I. Dale Ball has joined the inspection force of the Supervisor of Radio, 8th inspection district, Detroit, in the capacity of assistant radio inspector.

T. J. Bindner, formerly radio operator at station KJR, Seattle, is associated with the Masterphone Sound Equipment Corp., of Seattle.

Sidney Bloomenthal has joined the research department of the Radio Corporation of America in New York. He was, until recently, laboratory assistant at Ryerson Physical Laboratory, University of Chicago.

Robert E. Cain has become affiliated with RCA Photophone, Inc., at New York City. Mr. Cain left the engineering department of the Philco Radio Company at Philadelphia to assume his new connection.

E. C. Carlson, until recently laboratory assistant with Acoustic Products Manufacturing Corp., of Stamford, Conn., has become engineer with Sonora, Inc., of Buffalo, N. Y.

Joseph A. Davis is in the radio engineering department of the Temple Corporation at Clearing, Ill. He was formerly connected with the research department of the Automatic Electric Company.

J. D. Durkee, formerly officer in charge Radio Central, Navy Department, Washington, is now chief communications engineer of Universal Wireless Communication Company, Inc., at Chicago.

OBITUARY

With deep regret the Institute announces the death of

Edward Beech Craft

Mr. Craft was born in Cortland, Ohio, September 12, 1881. He received a common and high school education in Warren, Ohio. From 1900 to 1902 he was superintendent of the lamp department of the Warren Electrical and Supply Company. Shortly after that he joined the Western Electric Company at Chicago. In 1907 Mr. Craft moved to New York City to become development engineer with the Western Electric Company. By 1918 he had become assistant chief engineer in charge of all development work in the engineering department of the Western Electric Company. In 1922 he became chief engineer in charge of all research activities of Western Electric. With the incorporation of Bell Telephone Laboratories in 1925 Mr. Craft became its executive vice-president, and later a director of the corporation.

His contributions to the telephone and radio art through important inventions were numerous.

He was elected a Fellow in the Institute in April, 1920, and was a member of several other scientific and engineering societies.

His death occurred on August 20th, 1929.

GEOGRAPHICAL LOCATION OF MEMBERS ELECTED SEPTEMBER 4, 1929

Transferred to the Fellow grade

New York	New York City, c/o RCA Communications, Inc.	Latimer, Chester W.

Transferred to the Member grade

California	Palo Alto, Federal Telegraph Company	Suydam, Clinton H.
Dist. of Columbia	Washington, 3314 Newark St.	Mirick, C. B.
	Washington, c/o Federal Radio Commission	Segal, Paul M.
Illinois	Chicago, 6658 Loleta Ave.	Pfaff, Ernest R.
Pennsylvania	Wilkinsburg, 204 Avenue F	Landon, V. D.

Elected to the Member grade

California	San Francisco, Signal Office, Presidio of San Francisco	Winner, William Lane, Jr.
Dist. of Columbia	Washington, Dept. of Commerce, Radio Division	Barron, J. H., Jr.
Illinois	Alton, 1124 Washington St.	d'Adrian, A. L. Duval
	Chicago, 1400 Union Trust Bldg.	Caldwell, Louis G.
Massachusetts	Waban, 100 Devon Road	Replogle, Delbert E.
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	Manchester, British Broadcasting Corp.	Bird, R. J.
	Norbury, SW, 119 Headcorn Road	Shaw, A. C.
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	Wellington, G. P. O. Box 638	Thow, Keith H.
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	Shanghai, Nanyang University, c/o K. C. Chang	Hawng, Ju-tsu

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	Los Angeles, 1467 Temple St.	Richards, W. S.
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	Palo Alto, 551 Lytton Ave.	Wagener, Winfield G.
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	Weehawken, 62 Liberty St.	Pollack, Dale
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PART II

TECHNICAL PAPERS

Proceedings of the Institute of Radio Engineers
Volume 17, Number 10 *October, 1929*

STATIC AND MOTIONAL IMPEDANCE OF A MAGNETOSTRICTION RESONATOR*

BY

E. H. LANGE AND J. A. MYERS

(School of Engineering, The Johns Hopkins University, Baltimore, Md.)

Summary—*The equivalent series inductance and resistance of a long solenoid with nickel-steel bar has been investigated in relation to the excitation frequency for frequencies up to 14000 cycles per second. Beyond 400 cycles per second, after which only a negligible flux penetrates to the center of the bar, the inductance varies inversely with the square root of the frequency, and the resistance directly with the square root of the frequency. The product of the resistance and inductance is substantially constant over the frequency range; also, the power factor angle of the bar is substantially constant. The field for the higher frequencies is shown to be confined largely to the circumference of the bar, at which zone the greater part of the magnetostrictive force is produced. The results are discussed in relation to the theory of flux distribution in the bar.*

The effect of motion of the bar under the action of magnetostriction has been measured in terms of the motional impedance, and a circle diagram obtained. The ratio of motional resistance to reactance, in the vicinity of mechanical resonance of the bar, is of the order of magnitude of one to six thousand. The theory of total impedance, static and motional, is given, and the nature of the angular displacement of the resonant circle indicated.

Self-excitation by means of one 201A tube was found to be possible for the particular solenoid used in the measurements, by introducing in series with each half of the solenoid an air-core inductance to improve the low reactance-resistance ratio of the static impedance.

INTRODUCTION

THE electrical impedance of a solenoid containing a bar of magnetic substance may, for convenience, be considered as divided into two types; the volts per ampere in the solenoid due to the effective resistance and reactance, and the volts per ampere in the solenoid due to longitudinal vibration of the bar under the action of a periodic mechanical force, in this instance due to magnetostriction. The former impedance is called, in this paper, the static impedance, and the latter the motional impedance.

The two types can be considered separately, for, when motional impedance is present at all, the whole resonance phenomenon occurs for so small a frequency variation, due to the extreme sharpness of resonance, that the static impedance is substantially constant over this frequency range.

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* Dewey decimal classification: R210. Original manuscript received by the Institute, April 15, 1929. Revised copy received June 19, 1929.

The variation of the static impedance with frequency, or what is the same, the effective resistance and reactance with frequency, is primarily of interest in connection with the experimental examination of the flux distribution within the bar, and subsequent comparison with the theoretical relations for flux distribution. These relations are also important in determining the active annular zone within the bar over which the major part of the periodic mechanical force is operative, and the allied relation for the rotation of the motional impedance circle, due to the fact that the coil current and the resultant bar flux are not in phase. This latter relation requires consideration of the varying magnitude and phase of the magnetic induction within the bar for various distances from center to circumference of the bar.

Apart from the utility of the static impedance characteristic with varying frequency for the primary purpose stated, the relatively simple experimental procedure necessary for determining the effective depth of penetration of flux into a bar may be of interest as a possible method of examination of a magnetic substance, in bar form, with particular reference to changes in structure at or near the surface of the bar.

Mechanical oscillations in bars due to periodic magnetostriction induced by periodic magnetic intensities have been extensively investigated by Prof. G. W. Pierce,[1] who also examined the frequency stabilizing action of such bars when excited by means of a vacuum-tube oscillator. Further work on the motional impedance properties of various magnetic materials has been carried out by K. C. Black.[2] The change in length of bars of iron, nickel, and cobalt has been studied under static conditions, with varying amounts of magnetic intensity and tensile stress by several observers.[3] Recently, attention has been given to certain nickel-iron alloys (permalloys),[4] in which the negative magnetostriction of nickel is, to a certain extent, balanced against the positive magnetostriction which the iron possesses up to its saturation point. An explanation of the relative ease with which magnetism spreads through such alloys, for which the gross magnetostriction is practically zero, has been given by L. W. McKeehan[5] in terms of atomic magnetostriction.

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[1] G. W. Pierce, "Magnetostriction Oscillators," *Proc.* American Academy of Arts and Sciences, April, 1928.

[2] K. C. Black, "A Dynamic Study of Magnetostriction," *Proc.* American Academy of Arts and Sciences, April, 1928.

[3] See bibliography.

[4] O. E. Buckley and L. W. McKeehan, *Phys. Rev.*, Second Series, **26**, No. 2; August, 1925.

[5] L. W. McKeehan, *Phys. Rev.*, Second Series, **26**, No. 2, August, 1925.

DESCRIPTION OF APPARATUS

The apparatus used for both the static and motional impedance measurements is shown in Fig. 1. A solenoid extending somewhat beyond the ends of the bar was used for excitation purposes. There are two coils, each having 812 turns consisting of two layers of No. 22 enamelled copper wire, connected in series, the whole being practically continuous except for a small separation necessary for the nodal-point support of the bar. This type of support is of course essential for the investigation of motional impedance, and consists of two *V*-shaped jaws of moderately sharp edges which can be screwed together to clamp the bar and support it in the axis of the solenoid. In order to establish a steady magnetic flux within the bar a *U*-shaped permanent magnet was attached at each end of a steel bar, mounted as shown in Fig. 1, thus forming a closed magnetic circuit through the specimen bar. This arrangement was used because of its simplicity and also to obviate any coupled circuit which would be formed by a separate d-c excitation circuit and consequent reaction in the solenoid circuit under a-c conditions. As relatively small polarizations are needed, this arrangement proved entirely satisfactory for present purposes.

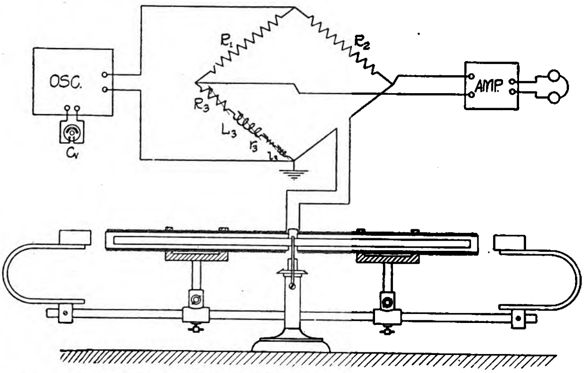

Fig. 1—Arrangement for measuring depth of flux penetration, static and motional impedance.

The bridge arrangement used for static and motional impedance measurements is also shown in Fig. 1. A type 8A Western Electric oscillator served as a driving source for the circuit. In measurements of the static impedance, the oscillator calibration was used over the frequency range, and for the frequency control in measurements of the motional impedance a vernier condenser, giving 2.5 cycles per sec. range, was used. The arms R_1, R_2, and R_3 are decade resistance boxes having non-inductive coils; L_3 is an 85-mh Brooks inductometer; l_3 is an inductometer for final balancing whose range is 0.6 mh; and r_3 is a 1-ohm slide wire. An amplifier circuit was connected to the arms as shown.

EXPERIMENTAL PROCEDURE

The oscillator was adjusted and allowed to warm up for at least four hours before making observations. This was found necessary in order to reduce frequency drift which, however, cannot be entirely eliminated. The slow frequency drift was of no consequence in the static impedance measurements. Observations were made to determine the resistance and reactance of the solenoid without the bar of magnetic material, for a frequency range of 100 to 14,000 cycles per second, also for the same range with the bar clamped in position in the axis of the coil. Sharp bridge balances were obtainable throughout this range.

The bar used in these experiments has its fundamental frequency at about 3834 cycles per second, and half of the resonance circle is traversed in less than one cycle. Owing to the extreme sharpness of resonance, any slow drift of oscillator frequency can add greatly to the difficulty of obtaining a sharp balance of the bridge for points on the motional impedance circle. After the oscillator drift was definitely identified this difficulty was easily obviated. It was found that by means of the vernier condenser any slow drift of frequency could be readily followed up, and sharp balances maintained anywhere on the motional impedance circle.

INDUCTANCE AND RESISTANCE CURVES

The measured inductance and resistance values give directly the series inductance and resistance of the solenoid. A typical curve of total inductance of solenoid with bar, L_i+L_0, for the frequency range 625 to 14,000 cycles per sec., is shown in Fig. 2. The curve L_0 represents the inductance due to flux linkages with the solenoid through the space outside the bar. This is obtained from the measured air-core inductance by computation on the basis of ratio of area of space surrounding the bar to total coil area. Due to the fact that it is not feasible for the motional impedance measurements to have a closely fitting coil, this procedure is necessary. Since the ratio of solenoid length to diameter is about 22 to 1, the computation of L_0 as indicated is sufficiently accurate. The difference between the total measured inductance and L_0 is the inductance due to flux linkages within the bar.

The curve of total inductance shows how rapidly the flux linkages per ampere in the coil are decreasing as the frequency is increased. From theoretical considerations, given in a following paragraph, (eq. 13 and 14), it appears that the effective flux-area of the bar, for frequencies sufficiently high to manifest skin-effect, is given by the relation $\pi a\delta$; in which a is the radius of the bar and δ the depth of penetration of the flux.

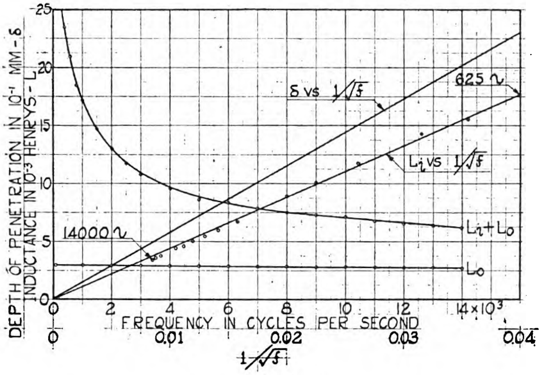

Fig. 2

Specifically,

$$\delta = 5033\sqrt{\frac{\rho}{\mu f}}\,\text{cm} \tag{1}$$

and the resultant flux vector is

$$\phi_i = \pi a\delta \cdot B_a(1-j)$$

in which B_a is the flux density at the surface of the bar. In accordance with equation (1) the inductance due to the bar was plotted against $1/\sqrt{f}$.

The linear relation between L_i and $1/\sqrt{f}$ holds very closely for the frequency range shown. Examination of this relation below 300 cycles has shown that as the frequency is decreased the straight line approaches a constant value of inductance, indicating an increasing uniformity of flux across the section of the bar. Further, the linear relation between L_i and $1/\sqrt{f}$ in the range shown indicates that μ is substantially constant in this range and for the excitation currents used (about 30 ma), so that from the measured value of ρ(90 x 10^{-6} ohms per cm cube) and experimental data on inductance variation with frequency, the depth of flux penetration δ can be computed. Thus

$$L_i = \frac{N\phi_i'}{I} \times 10^{-8} = 4\pi N^2 \left(\frac{\text{effective area}}{\text{length}}\right) \times \mu \times 10^{-9} \cdots \text{henries.}$$

Substituting the value of the effective area in terms of the depth of penetration, and noting that ϕ_i' in the above equation is the component of the bar flux in phase with the coil current,

$$L_i = \left(\frac{5033 \times 4\pi^2 \times 10^{-9} N^2 \cdot a \cdot \sqrt{\rho\mu}}{\text{length}}\right) \cdot \frac{1}{\sqrt{f}} \tag{2}$$

The value of μ computed from the experimental curve shown, using equation (2) is $\mu = 64.9$. Thus, for the bar under test the depth of flux penetration as related to the frequency is

$$\delta = \frac{5.91}{\sqrt{f}} \text{ cm} \tag{3}$$

Length of bar = 58.45 cm
Radius of bar = 0.635 cm = a
Mean diameter of coil = 2.70 cm

Analysis of the resistance of the solenoid is somewhat more involved. The curve of resistance, R_c with frequency, for the coil without the bar is shown in Fig. 3. The total resistance of the solenoid and bar is indicated by R_t, and the difference $R_t - R_c$ is the series resistance due to the bar alone. The experimental values of $(R_t - R_c)$ plotted against $\sqrt{f}$ show a slight curvature up to several hundred cycles per second, until the flux has ceased to penetrate to the center of the bar, after which the resistance is substantially linear with $\sqrt{f}$

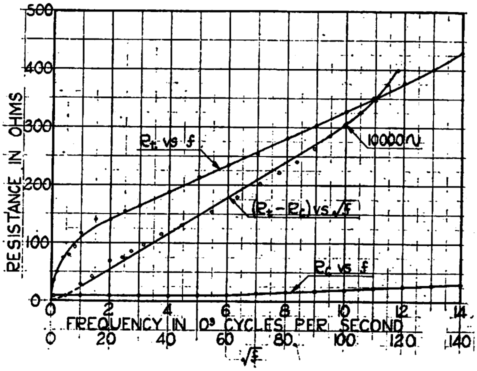

Fig. 3

over the whole range. The deviation from linearity in the vicinity of 10,000 cycles is believed to be due to the fact that at the higher frequencies the true resistance due to the bar cannot be accurately obtained by deducting the coil resistance, for the field distribution about the coil with the bar removed is somewhat changed.

The equivalent series resistance due to the losses in the bar, i.e. $(R_t - R_c)$ is made up of eddy-current and hysteresis effects. The short circuit or circulating currents around the bar are confined largely to the outer circumference for frequencies above 1000 cycles, and equation (1) measures both the effective depth of penetration of the flux and short-circuit current. The power loss due to the bar in terms of the coil current I is

$$P = I^2(R_t - R_c) = I^2(R_e + R_h) \tag{4}$$

in which R_e is the series resistance due to eddy currents and R_h due to hysteresis. In the case of R_e, since the short-circuit currents are proportional to the current I, and the effective section of the short-circuit path directly proportional to δ, the value of R_e will vary directly with $\sqrt{f}$. In the case of R_h, the effective volume is proportional to δ, i.e., $1/\sqrt{f}$, so that as far as effective volume and number of hysteretic cycles are concerned, R_h varies directly as $\sqrt{f}$; the lack of complete linearity between I^2 and the energy loss per loop in its relation to some exponent of the flux density is, to a large extent, compensated in the total loss by the fact that the coil current I does not vary over a wide range, also by the considerable excess of eddy-current loss over hysteresis loss. The relations for the variation of R_e and R_h with frequency can be more clearly seen from the following equations. The power loss due to the eddy currents is

$$I^2R_e = I_b^2R_b = \frac{2\pi a\rho}{l\delta}\cdot I_b^2\cdot$$

Thus,

$$R_e = \left(\frac{2\pi a\rho}{l}\,\frac{I_b^2}{I^2}\right)\cdot\frac{1}{\delta}$$

where I_b is the aggregate circulating current in the bar, l the bar length, a the bar radius, and R_b the resistance for the circulating current. Taking into account the actual distribution of current across the section of the bar, the aggregate current I_b can be shown to be equal to the effective current density at the surface of the bar $u_a/\sqrt{2}$ multiplied by $\delta \times l$, the effective cross section; also the current I_b to be directly proportional to the surface value of flux density B_a, i.e., to the coil current, and finally, the resistance R_b to be that of a hollow tube of thickness δ, radius a and length l. The power loss due to hysteresis can be expressed in the form $I^2R_h = \text{constant} \times \text{vol.} \times f \times B^n$, and as the effective volume of material in which the hysteresis loss is taking place is proportional to δ, this becomes

$$R_h = \text{constant} \times \sqrt{f} \times \frac{B^n}{I^2}$$

THEORY OF INDUCTANCE VARIATION

The distribution of a harmonically changing flux in a solid conductor has been treated by numerous writers.[6] In the particular problem at hand what is needed is not only the flux distribution, but also the resultant flux in phase and magnitude, in order that the distribution may be determined in terms of its manifestations in the solenoid surrounding the bar. The bar can be considered to be divided into annular zones, and the vector sum of the fluxes obtained, also the phase of the resultant flux with reference to the coil current. For present purposes it is desirable to limit the discussion to the distribution of flux for frequencies sufficiently high to reduce the field in the central portion of the bar to negligible values. Only such frequencies are of interest here, and the equations under these conditions are very much simplified.

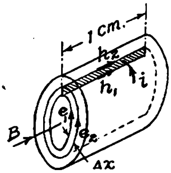

Fig. 4

Referring to Fig. 4, the voltage difference between two successive cylinders due to changing magnetic induction is

$$e_2 - e_1 = -2\pi x \Delta x \frac{\delta B}{\delta t} \times 10^{-8}.$$

Thus we may write

$$\frac{\delta F}{\delta x} = -\frac{\delta B}{\delta t} \times 10^{-8} \tag{5}$$

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[6] Alex. Russell, "Alternating Currents," Vol. I; J. H. Jeans, "Electricity and Magnetism."

in which F is the electric intensity along the circumference.[7] For any area of elementary width parallel to the induction, the m.m.f. of the circulating current is

$$h_2-h_1=-0.4\pi u\Delta x,\ \text{or}\ \frac{\delta h}{\delta x}=\frac{1}{\mu}\ \frac{\delta B}{\delta x}=-0.4\pi u \tag{6}$$

Also, the current around any ring in terms of the resistivity ρ (ohms per cm cube) and the current density u (amperes per cm^2) is

$$e/2\pi x\rho/\Delta x=u\Delta x \quad \text{or}\ F=\rho u \tag{7}$$

Combining (5, 6, 7)

$$\frac{\delta^2 B}{\delta x^2}=\frac{4\pi\mu\times 10^{-9}}{\rho}\cdot\ \frac{\delta B}{\delta t}=k\frac{\delta B}{\delta t} \tag{8}$$

and

$$\frac{\delta^2 u}{\delta x^2}=k\frac{\delta u}{\delta t}\cdot$$

Solutions of (8) are of the form $B=B_m\epsilon^{j\omega t}\cdot\epsilon^{(\alpha+j\beta)x}$ where

$$\alpha+j\beta=\pm\sqrt{2}j\sqrt{\frac{k\omega}{2}}=\pm(1+j)\sqrt{\frac{k\omega}{2}}\cdot$$

Put

$$\gamma=\sqrt{\frac{k\omega}{2}};\ \omega=2\pi f.$$

Then,

$$B=\left[B_1\epsilon^{(1+j)\gamma x}+B_2\epsilon^{-(1+j)\gamma x}\right]\epsilon^{j\omega t} \tag{9}$$

Noting that the flux density must have circular symmetry about the axis of the bar, $B_1=B_2=B_m$, and that for the condition assumed of negligible flux density in the central portion of the bar in comparison with the flux density at the surface, the second term of (9) may be neglected, the equation becomes

$$B=B_m\epsilon^{j\omega t}\cdot\epsilon^{(1+j)\gamma x} \tag{10}$$

Substituting $x=a$ to obtain the flux density at the circumference,

$$B_a=B_m\epsilon^{j\omega t}\cdot\epsilon^{(1+j)\gamma a} \tag{11}$$

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[7] Strictly, $\partial(Fx)/\partial x=-x(\partial B/\partial t)\times 10^{-8}$; however, under the conditions above assumed, (5) is sufficiently accurate.

The vector flux density in terms of its value at the circumference is therefore

$$B = B_a \epsilon^{(1+j)\gamma(x-a)}.$$

If, now, distances are measured from the circumference instead of from the center, $z = a - x$, and the vector flux density is

$$B = B_a \epsilon^{-(1+j)\gamma z}. \tag{12}$$

The instantaneous value is therefore $B = B_a \epsilon^{-\gamma z} \cos (\omega t - \gamma z)$. From (12) γ is a reciprocal length. Put $\delta = 1/\gamma$, then

$$\delta = \frac{10^4}{2\pi}\sqrt{\frac{10\rho}{\mu f}} = 5033\sqrt{\frac{\rho}{\mu f}} \text{ cm}. \tag{13}$$

The phase of the flux varies continuously from one annular zone to the next, so that the resultant bar flux is the vector sum of the fluxes of the various annular zones. The flux of any zone of radius x is

$$d\phi = 2\pi x dx \cdot B = -2\pi(a - z)Bdz$$

and the total flux between the circumference and any inner zone is

$$\phi = 2\pi B_a \int_0^z (z - a)\epsilon^{-\gamma(1+j)z} \cdot dz.$$

Evaluation for the whole bar gives

$$\phi = 2\pi a \frac{\delta}{2} B_a(1 - j). \tag{14}$$

Equation (14) shows that the magnitude of the flux components is obtained by considering the surface value of density B_a as spread uniformly over an effective area equal to the circumference $2\pi a$ times one half the depth of penetration δ. The phase of the resultant flux is 45 deg., lagging the surface density B_a which is in phase with the coil current.

The above result shows that, provided the frequencies are such as to confine the flux largely to the outer rim, the ratio of the reactance to eddy-current resistance is constant for all frequencies, and equal to one. The constancy of the ratio is also shown by the experimental curves, for $L\sqrt{f}$ is constant, also $R/\sqrt{f}$ is constant, so that by division fL/R is constant; i.e., the eddy-current angle is independent of frequency. The experimental curve for resistance includes the hysteresis resistance, so that in this case the reactance-resistance ratio is constant and less than one, that is, the resultant bar flux lags the current in the coil by an angle greater than 45 deg. It is of interest to note that because of the skin effect, which gives rise to the particular relations above stated between L, R, and $\sqrt{f}$, the product of the inductance L and resistance R is constant and independent of frequency. Attention appears to have been first directed to the constancy of this product for the magnetic circuit of telephone receivers, by means of an empirical relation.[8]

Taking the coil current vector for reference, and denoting by θ_i the total angle of flux lag (eddy-current and hysteresis), the flux vector is $\phi_i[\cos\theta_i - j\sin\theta_i] = \phi_i\cos\theta_i(1-jp)$ in which ϕ_i is the total flux, and $p=\tan\theta_i > 1$. The coil e.m.f. necessary to maintain the flux ϕ_i is therefore

$$j\omega N\phi_i\cos\theta_i(1-jp).$$

Substituting for $\phi_i\cos\theta_i$ its value in terms of the inductance and coil current, the e.m.f. becomes

$$j\omega LI(1-jp)=\omega LI(p+j).$$

Thus the impedance due to the bar is

$$Z_s=\omega L(p+j). \qquad (15)$$

The copper coil resistance R_c and leakage inductance L_0 give rise to the impedance $Z_c=R_c+j\omega L_0$. The total impedance when the bar is not in motion is therefore the vector sum $Z=Z_c+Z_s$.

MOTIONAL IMPEDANCE CIRCLE

A typical motional impedance circle for the nickel-steel bar used in these experiments is shown in Fig. 5. The line Z_s is the static impedance line for the frequencies shown, and includes the copper coil resistance R_c and the leakage reactance due to L_0. The maximum room temperature variation during the measurements was less than one-half deg. C. Between the frequency 3750 and 3900, the resistance and reactance values follow around the circle shown, with frequency increasing in the clockwise direction. The origin of the motional reactance and resistance axes is indicated by the intersection of the static impedance line and the motional impedance circle. In accordance with (23) the ratio of motional resistance to motional reactance is 1 to 6500. The small energy loss of the bar also manifests itself in the persistence of oscillations which beat with the driving source, when the impressed frequency is changed so as to pass through the bar frequency. For the frequencies of the motional circle, the bar emits sound of considerable intensity, which becomes very penetrating and disagreeable when listened to for a considerable length of time. A part of the mechanical resistance is sound radiation resistance, so that the ratio of actual mechanical resistance due to internal losses of the bar, to the reactance, is even less than the above value. The reactance and resistance variations with frequency, in the vicinity of resonance, are shown in Fig. 6 and Fig. 7.

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[8] A. E. Kennelly and G. W. Pierce, "The Impedance of Telephone Receivers as Affected by the Motion of Their Diaphragms," *Electrical World*, September 14, 1912.

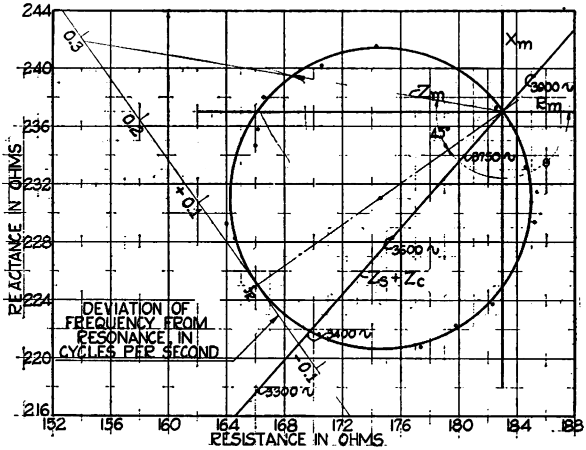

Fig. 5—Static and motional impedance curves.

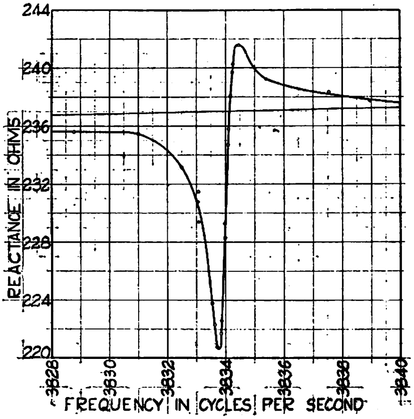

Fig. 6

Motional Impedance Theory

The characteristic magnetostriction curves for iron under a constant tension indicate that as the magnetic intensity is increased the deformation is positive up to a certain value of H, after which the deformation decreases and becomes negative. A family of such curves results for different constant values of tension. In the present arrangement, since a steady polarizing field is used with a small superposed alternating field, the increment of the main field is of principal interest in relation to any change of tension or deformation. Calling the tension p, and the deformation or strain q, the increment of the flux density may be stated in the form

$$dB = K_p p + K_q q$$

where $\delta B/\delta p = K_p$, and $\delta B/\delta q = K_q$. The sign of the coefficients is determined from the characteristic curves for the particular magnetic material and polarizing field. Thus,

$$dB = dB_i + dB_q \tag{16}$$

in which dB_i may be regarded as the incremental change due to the external alternating field which produces a tension dB_i/K_p, and $dB_q = K_q q$ the change due to increase or decrease in the length of the bar. The time rate of change of dB_i and dB_q gives rise to the ordinary voltage and the motional voltage, i.e., for any harmonic coil current, to the ordinary impedance Z_s and the motional impedance Z_m.

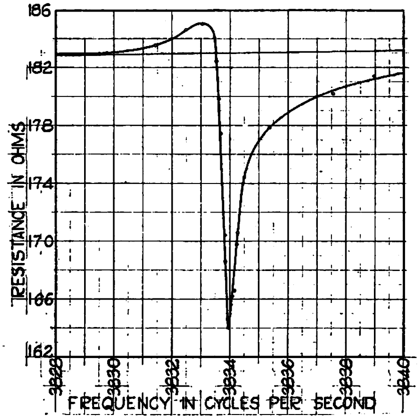

Fig. 7

Under ordinary conditions, the deformation is so small as to cause a negligible change in the flux, but when the excitation frequency is such as to cause mechanical resonance of the bar the deformation is greatly increased, and the motional impedance becomes a part of the total impedance. Computation of the motional impedance caused by

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the solid bar is complicated by the fact that the flux distribution is not uniform across the section of the bar and the deformation is not uniform along the bar; however, the motional impedance in terms of basic factors can be closely approximated by taking account of the simplest aspect of these effects.

The effect of flux distribution will be considered first, with reference to the ordinary flux and the deformation flux. The force at any one of the annular zones previously considered is $\Delta P = 2\pi x \Delta x \cdot B_i / K_p$, and in accordance with the development of equations (14) and (15) the total force or vector sum of all the forces at any section is

$$P = \frac{\phi_i}{K_p} \cdot \epsilon^{-j\theta_i} = \frac{\phi_i}{K_p} \cos \theta_i (1 - jp).$$

The resultant force vector therefore lags the current in the coil by the angle θ_i. Substituting $\phi_i \cos \theta_i = LI/N$.

$$P = \frac{LI}{K_p N \cos \theta_i} \epsilon^{-j\theta_i}. \tag{17}$$

The sinusoidal force P gives rise to a sinusoidal deformation q, and to a resultant deformation flux ϕ_q. The relation $B_q = K_q \cdot q$ for the deformation flux density in terms of the deformation is true for static deformations or for a very fine wire in which the deformation is rapidly alternating; for a solid bar mechanically excited at high frequencies the flux distribution is governed by (8). The deformation flux as influenced by the eddy currents is assumed in accordance with (14) to be $\phi_q = \pi a \delta \cdot B_{qa} \cdot (1 - j)$, and since the deformation flux is also subject to the effects of hysteresis, this may be written $\phi_q = \pi a \delta \cdot B_{qa}(1 - pj)$, where B_{qa} is the deformation flux density at the surface of the bar and $p' = \tan \theta_q > 1$, θ_q being the angle by which the resultant deformation flux lags the surface flux density B_{qa}. The surface zone of infinitesimal thickness is free from the counter m.m.f. of the interior circulating currents, and it may be assumed that for this zone $B_{qa} = K_q \cdot q$, thus

$$\phi_q = \pi a \delta K_q q (1 - jp') = \frac{K_q A \epsilon^{-j\theta_q}}{\cos \theta_q} \cdot q \tag{18}$$

where A is the effective flux area.

The effect of variation of the deformation q along the bar will now be considered. When the bar is excited at its fundamental resonant frequency, a stationary wave of deformation exists along the bar, the deformation being a maximum in the center and zero at the ends.

The extreme sharpness of resonance indicates that the energy loss per cycle per unit length of bar is very small compared with the stored energy per unit length, so that the system closely approximates a purely reactive resonator, and the deformation may be considered as sinusoidal in distribution along the bar, all points being in the same time phase. In accordance with (18) the deformation flux is therefore distributed sinusoidally along the bar, and on the assumption that all of the deformation flux links with the turns of the solenoid, the total flux linkages are $(2/\pi)N\phi_q$. Denoting by N_q the effective turns of the solenoid for the deformation flux linkages, the total linkages will in general be $N_q\phi_q$. The e.m.f. which the line must supply to overcome the motional e.m.f. of the bar in the coil is therefore

$$E_m = N_q\dot{\phi}_q = \frac{N_qK_qA\epsilon^{-j\theta_q}}{\cos\theta_q}\cdot\dot{q}_0, \tag{19}$$

in which $\dot{q}_0$ refers to the time rate of change of the deformation halfway between the ends of the bar.

The magnitude and phase of $\dot{q}_0$ in relation to the total exciting force P are determined by the mechanical resonance properties of the bar. The mechanical properties of the resonator for the frequencies at or near resonance, can in the present case of a highly reactive system be most simply expressed in terms of an equivalent lumped resonator. Denoting by m the mass of the resonator, r the resistance, g the elastic constant, l the length of the resonator, A_b the cross-section, q the deformation, and y the displacement of the end of the resonator, then for a lumped resonator

$$\frac{P}{l} = m\ddot{q} + r\dot{q} + gq.$$

When the total exciting force P is a harmonic function of the time, the rate of change of deformation is

$$\dot{q} = \frac{P}{[r + j(\omega m - g/\omega)]l} \tag{20}$$

in which $r + j(\omega m - g/\omega)$ is the mechanical impedance of the resonator. The phase angle of $\dot{q}$ with reference to the force P is

$$\psi = \tan^{-1}\frac{\omega m - g/\omega}{r}. \tag{21}$$

and mechanical resonance occurs for the frequency given by

$$\omega_0=\sqrt{\frac{g}{m}}. \tag{22}$$

In the actual resonator, the quantities, $\ddot{q}$, $\dot{q}$, and q, are sinusoidally distributed along the bar, and for any one of these quantities the values along the bar are in time phase. The actual resonator thus corresponds to numerous lumped resonators running in synchronism but having different amplitudes. The total energy content of the actual resonator, i.e., the kinetic and potential energy, can be readily determined in terms of the maximum velocity V and displacement Y at the ends of the bar, and the mass m and elastic constant g determined in relation to the actual resonator. Since the velocity distribution is sinusoidal, the mean square velocity is $V^2/2$, thus if M is the mass of the actual resonator the equality of kinetic energies gives $(m/2)V^2=M/2\cdot V^2/2$, or $m=M/2$. The equivalent lumped mass is therefore one-half the actual mass. Equality of the potential energies of the lumped and actual resonators gives (average $q^2e/2)A_bl=gY^2/2$, in which e is Young's modulus and $q^2e/2$ is the potential energy per unit volume. Noting that the maximum displacement of the ends of actual resonator is $Y=q_{\text{av}}\cdot l/2=(2/\pi)q_0\cdot l/2$, and that the mean square deformation is $q_0{}^2/2$, the equivalent elastic constant reduces to $g=\pi^2/2\cdot A_be/l$. Thus the equivalent elastic constant is $\pi^2/2$ times the value for a uniformly stressed bar.

The mechanical resistance can be compared with the reactance by determining the frequency deviation from resonance necessary to make $\psi=45$ deg. Denoting by ω_1 the deviated angular velocity, and noting that due to extreme sharpness of resonance $\omega_0+\omega_1$ may be replaced by $2\omega_0$,

$$\frac{\omega_1 m-g/\omega_1}{r}=1=\frac{(\omega_1{}^2-\omega_0{}^2)m}{\omega_1 r}$$

or

$$\frac{r}{\omega_0 m}=\frac{2(f_1-f_0)}{f_0}=\frac{r}{\sqrt{gm}}. \tag{23}$$

The motional e.m.f. given by (19) in terms of the maximum rate of change of deformation can now be expressed in terms of the force by means of (17) and (20), and the motional impedance obtained. Thus

$$E_m=\frac{\pi N_qK_qAP\epsilon^{-j\theta_q}}{\cos\theta_q\cdot l[r+j(\omega m-g/\omega)]}=\frac{\pi N_qK_qALI\epsilon^{-j(\theta_i+\theta_q)}}{K_pNl\cos\theta_i\cos\theta_q[r+j(\omega m-g/\omega)]}$$

or

$$Z_m=\frac{E_m}{I}=\frac{K\epsilon^{-j(\theta_i+\theta_q)}}{r+j(\omega m-g/\omega)}. \tag{24}$$

The coefficient K is used to designate the product of the coefficients in the previous equation. The reciprocal of the mechanical impedance gives a circle locus with maximum electrical motional impedance along the resistance axis; however, due to eddy currents and hysteresis this maximum has an angular displacement which causes the motional e.m.f. to lag behind the coil current. The area A in (24) is the effective flux area as defined in (14) and is small compared with the area of the bar. In the case of a tube in which the thickness of the walls is less than δ, the area A approaches in value the actual cross-section of the tube, and the angles θ_i and θ_q are decreased, thereby decreasing the angle of lag between the coil current and the motional e.m.f.

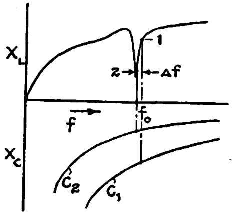

Fig. 8—Frequency stabilizing action of motional inductance.

VECTOR DIAGRAM OF RESONATOR

Referring to Fig. 9, the results of the foregoing equations are shown in the form of a vector diagram. The ordinary flux ϕ_i lags the coil current I by the angle θ_i which is substantially independent of frequency. The part of the impressed e.m.f. necessary to maintain this flux is shown by E_s, 90 deg. in advance of the flux, and E_c is the part of the impressed e.m.f. necessary to overcome the leakage reactance and resistance of the solenoid. When the motion of the bar is negligible, the vector sum of E_s and E_c or E gives the total impressed e.m.f. necessary to maintain the current I. Since the resonance of the bar is

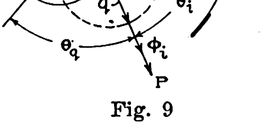

Fig. 9

very sharp, the frequency may be considered variable over a sufficient range to include resonance without appreciably changing the phase or magnitude of the vectors E_s or E_c. The force P is in phase with the flux ϕ_i and at exact resonance the rate of change of deformation $\dot{q}$ is in phase with P. The dotted circle shows the locus of $\dot{q}$ as the frequency is changed. The motional e.m.f. E_m lags the rate of change of deformation by the angle θ_q; the resonance circle for the motional e.m.f. is shown by the solid line. The total impressed e.m.f. is the vector sum of E_s, E_c, and E_m, or the total impedance the vector sum of Z_s, Z_c, and Z_m, the impedances being given by (15) and (24).

USE AS A FREQUENCY STABILIZER

The action of the bar in stabilizing frequency, that is, in rendering the generated frequency to a large extent independent of the capacity and static inductance of a vacuum-tube oscillator circuit, may be seen by reference to Fig. 8, which shows the reactance diagrams for variable frequency of a circuit having static and motional inductance, and capacitance. The upper curve shows the static and motional inductive reactance, and the lower curves are the capacitive reactance for two different constant values of capacitance. For frequencies in the vicinity of mechanical resonance of the bar, for a variation of Δf, there are a group of inductance values between the points 1 and 2 available for balancing any capacitive reactance in the range C_1 to C_2. Thus, a relatively large variation of capacitance C produces a small change of frequency between zero and Δf. The motional impedance vector simply shifts to a new position on the resonance circle, the whole of which is executed in an extremely small frequency variation. The extent of frequency stabilization depends therefore upon the sharpness of mechanical resonance $\Delta f/f_0$ and upon the magnetostriction activity of the bar, that is, upon the extent to which the kinetic and potential energies of the oscillatory system can be concentrated in the mechanical rather than the electrical system. The larger the motional impedance circle in relation to the static impedance value at the resonant frequency, the greater may the variation of capacitance be without appreciable frequency variation.

The arrangement of coil and bar shown in Fig. 1 was found to operate satisfactorily in a Hartley circuit, using a 201A tube, and additional inductance to increase the ratio of reactance to resistance, in both the grid and plate branch of the solenoid. This of course further decreases the motional inductance in relation to the total inductance; however, even under these conditions, the stabilizing action of the bar is evident for a variation of the capacitance connected between the grid and the plate.

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CALCULATION OF CHARACTERISTICS AND THE DESIGN OF TRIODES*

BY

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Summary—*The primary object of the paper is to present a simple method of designing triode vacuum tubes. It comprises three parts; the first part deals with the calculation of characteristics and constants of the triode from the electrode structures, special considerations being taken for applying the formulas already established for tubes of complicated structures. In the second part the derivation of various working conditions of a triode from its static characteristic is treated, in which the writer worked out a graphical representation of d-c and a-c components of the anode working current at various working voltages. The resulting dynamic characteristic diagram is applicable to any type of triodes in evaluating the working voltages, currents and powers, whether the tube be used as an amplifier, oscillator, or modulator. The third part presents the designing procedure for a typical case in which the use of the triode is indicated and its power output is given. First the working points are determined on the dynamic characteristic diagram so as to give the required working condition; then all the quantities arising in the operation are known in their relative amounts. The type of the tube is selected between the high-impedance type and the low-impedance one, which necessitates technical and economical considerations. Then all the quantities are known in numerical values and characteristics of the tube are determined. The electrode structures are calculated to give the required characteristics.*

The present paper, being an abridgment of a paper published in Japan,[1] *is primarily intended for presenting a simple method of designing triodes.*

Part 1: Computation of Characteristics from Electrode Structures

THE well-known space-charge equation[2]

$$i_a = Ge_a^{3/2} \tag{1}$$

applies to diodes in actual cases, when it has a cathode of uniform potential all over the surface or when the anode voltage is very high compared with the filament voltage. This equation may also be applied, in ordinary cases, if the anode voltage is referred to the neutral point of the cathode, or when the filament is lighted by alternating current.

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* Dewey decimal classification: R 130. Original manuscript received by the Institute, April 9, 1929.

[1] Y. Kusunose, "Calculation on Vacuum Tubes and the Design of Triodes," Researches of the Electrotechnical Laboratory, No. 237; September, 1928, (written in English). Published by Kōseikai, No. 1/1 Yūrakucho, Tokyo, Japan. (Price about $1.70).

[2] I. Langmuir, *Phys. Rev.*, 2, 450, 1913.

When the cathode consists of a filament heated by direct current at a terminal voltage e_f, the above equation must be corrected for the non-uniformity of anode-to-cathode potential difference along the length of the filament, and the modified forms[3] are as follows:
for $e_a < e_f$:

$$i_a = \tfrac{2}{5} G \frac{1}{e_f} e_a^{5/2} = \tfrac{2}{5} G e_f^{3/2} \left(\frac{e_a}{e_f}\right)^{5/2} \tag{2}$$

and for $e_a > e_f$:

$$i_a = \tfrac{2}{5} G \frac{1}{e_f} \left[e_a^{5/2} - (e_a - e_f)^{5/2}\right] = \tfrac{2}{5} G e_f^{3/2} \left[\left(\frac{e_a}{e_f}\right)^{5/2} - \left(\frac{e_a}{e_f} - 1\right)^{5/2}\right] \tag{3}$$

The two equations may be reduced to a single one, as

$$i_a = \tfrac{2}{5} G e_f^{3/2} \cdot f\left(\frac{e_a}{e_f}\right). \tag{4}$$

The constant G which is to be called "perveance" is determined from the electrode configurations, and in case of cylindrical anode,

$$G = 2.33 \times 10^{-6} \frac{A}{x_a^2} \tag{5}$$

in which A = effective anode area; x_a = distance of the anode surface from the axis of the cathode, or radius of the anode in this case.

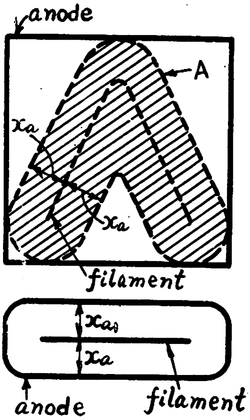

Fig. 1—Effective area of plane anode.

All the quantities relating to dimensions are to be expressed in centimeter units, and those relating to electricity in volts, amperes, ohms, watts, etc., unless otherwise specified.

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[3] Van der Bijl, "Thermionic Vacuum Tube," p. 64, 1920.

The writer has found that the above expression of G is equally applicable to anodes of plane forms, if the effective anode area is taken as that area comprised in the breadth $2x_a$ along the filament length projected on the anode, as illustrated in Fig. 1.*

TRIODE CHARACTERISTICS

In a triode, an anode voltage e_a and a grid voltage e_g produce as their combined effect a certain strength of electric field around the cathode, and the same electric field can be produced if a single anode exists at the grid suface and is at the potential[4]

$$e_g' = \frac{e_a + k e_g}{1+k} \tag{6}$$

where k is the amplification constant.

The electron current is mainly governed by the electrostatic action of the system of electrodes, and if it be assumed that the two cases above cited give equal electron current, the characteristic of the triode

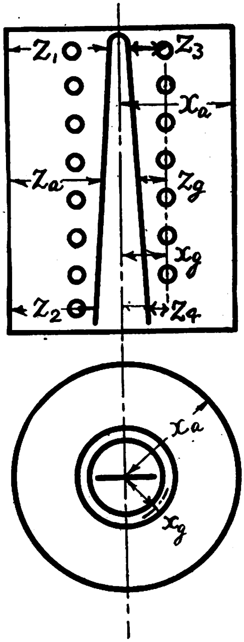

Fig. 2—$Z_a = (Z_1 + Z_2)/2$
$Z_g = (Z_3 + Z_4)/2$

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* This has lately been proved theoretically by N. Kato and S. Koizumi.

[4] W. H. Eccles, "Continuous Wave Wireless Telegraphy," Part I, p. 338; 1921.

will be expressed by the following equations which are obtained by replacing e_a in (1) and (4) by e_g':

$$i = Ge_g'^{3/2} \tag{7}$$

$$i = \tfrac{2}{5}Ge_f^{3/2}\cdot f\left(\frac{e_g'}{e_f}\right) \tag{8}$$

in which $f(e_g'/e_f)$ may be obtained from Table I.

TABLE I

e_g'/e_f	$f(e_g'/e_f)$
0	0
0.25	0.031
0.50	0.177
0.75	0.414
1.00	1.00
1.5	2.57
2.0	4.65
2.5	7.13
3.0	9.94
4.0	16.51
5.0	24.1
6.0	32.5
8.0	53.0
10	74.0
15	138
20	211
40	622

In this case

$$G = 2.33\times 10^{-6}\frac{\text{grid surface area}}{x_g^2}$$

and for cylindrical electrodes, grid surface area $=2\pi x_g l$ and anode area $=A=2\pi x_a l$, l being the effective length of the electrodes.

Hence

$$G = 2.33\times 10^{-6}\frac{A}{x_a x_g}$$

in which x_a and x_g are the distances of the anode and grid surfaces respectively from the cathode axis. But the writer found by experiment that in case of triodes having cylindrical anode and grid, and a cathode of other than a single axially-spanned filament, such as V-shaped, the values of x_a and x_g in the expression of G should be replaced by the mean shortest distances from the respective electrodes to the cathode z_a and z_g which should be taken in such a way as illustrated in Fig. 2.†

Then generally

$$G = 2.33\times 10^{-6}\frac{A}{z_a z_g}. \tag{9}$$

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† This can be verified by conformal transformation and integration as worked out by S. Koizumi.

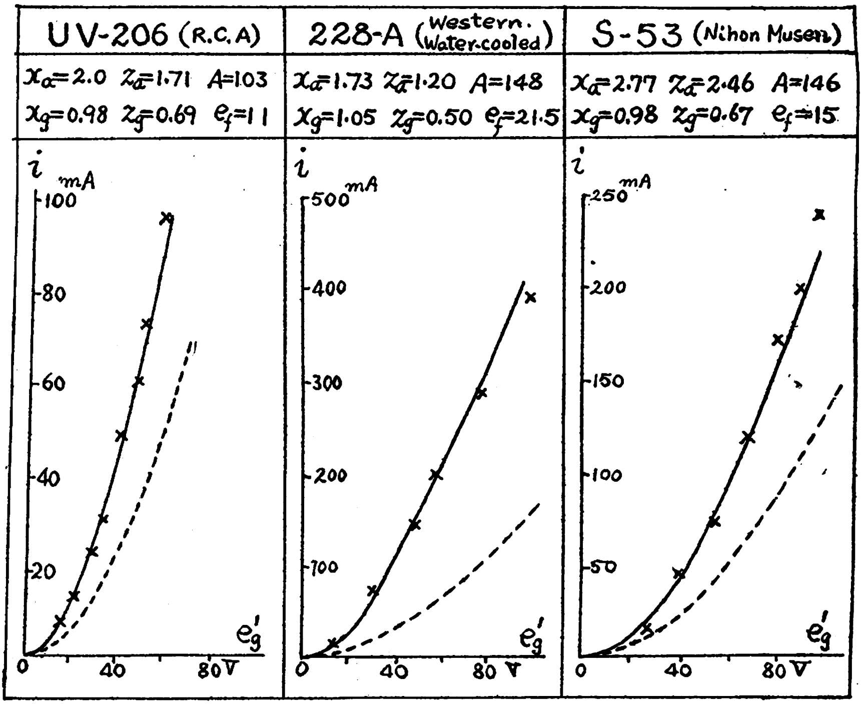

Fig. 3—Calculated characteristics of tubes with cylindrical anode and *V*-shaped filament.

Calculated with $G=2.33\times10^{-6}$ $x\dfrac{A}{Z_aZ_g}$ ————————

Observed points x. $x\dfrac{A}{X_aX_g}$ — — — — —

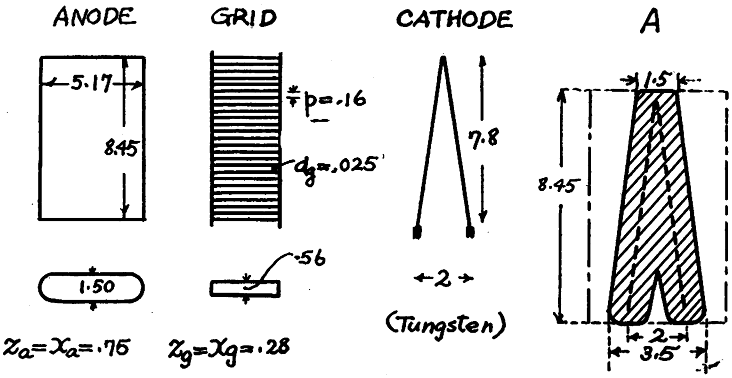
ANODE
GRID
CATHODE
A
5.17
8.45
p=.16
d_g=.025
7.8
2
(Tungsten)
1.50
.56
$Z_a=X_a=.75$
$Z_g=X_g=.28$
1.5
8.45
2
3.5

Fig. 4—Electrode dimensions and effective anode area of type UV204 tube.

This expression may also be applied to tubes with plane forms of anode and grid and in this case evaluation of the effective anode area A should be made as described before (Fig. 1).

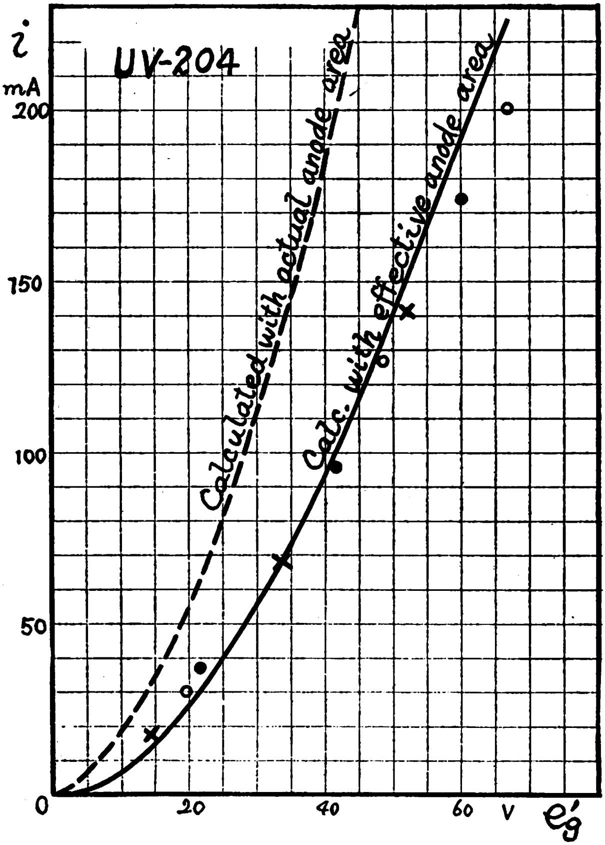

Fig. 5—Calculated characteristics of UV204.

$$\text{Observed points}\begin{cases} x \text{ at } e_a = 1500 \text{ v} \\ \bullet \quad " \quad " = 2000 \; " \\ 0 \quad " \quad " = 2500 \; " \end{cases}$$

Example 1: Calculation of characteristics of triodes by equation (8), one with the value of G obtained from expression (9) (full lines in Fig. 3), and the other with $G=2.33\times10^{-6}\ A/x_ax_g$ (dotted lines in Fig. 3). Observed characteristics (points in Fig. 3) show the validity of taking the former expression of G.

Example 2: Evaluation of the effective anode area of a plane anode. A triode type UV204 (manufacturer: Tokyo Denki Co.) has dimensions as shown in Fig. 4. The effective anode area obtained after the principle described above is $A=2\times(8.45\times(1.5+3.5)/2)=42.3$ and for this area $G=0.446\times10^{-3}$. If, on the other hand, actual area of the anode surface is

taken $A'=2\times(8.45\times5.17)=87.3$ and for this area $G'=0.965\times10^{-3}$. The characteristics are calculated by equation (8) for both and are plotted in Fig. 5. Observed data shows the validity of taking $A=42.3$.

The depiction of the anode-current grid-voltage characteristic curves may be simplified in the following manner. First obtain the numerical relation between i and (e_g'/e_f), and from it the relation of i to e_g' for a given value of e_f. Then assume $e_a=0$, or $e_g'=\{k/(1+k)\}e_g$, and i versus e_g curve can be drawn and this corresponds to the characteristic for the anode voltage zero. In order to obtain a characteristic for any anode voltage e_a, this curve for $e_a=0$ should be shifted horizontally toward the negative direction of the grid voltage by an amount of e_a/k in the scale of grid voltage.

It should be remembered that a slightest error in the calculated value of k causes a considerable amount of error on the resulting values of i, especially at high anode voltages.

Example 3: Computation of characteristics of a triode type 211A (Western Electric Co.). Dimensions are shown in Fig. 6.

Calculations:

$$A=2\times\left(5.7\times\frac{2.9+3.1}{2}\right)=34.2$$

$$Z_a=x_a=0.68 \qquad Z_g=x_g=0.25$$

$$G=2.33\times\frac{34.2}{0.25\times0.68}\times10^{-6}=0.469\times10^{-}$$

$$e_f=9.0 \qquad 2/5Ge_f^{3/2}=0.0051$$

$$i=0.0051f\left(\frac{e_g'}{e_f}\right).$$

Characteristic curve representing the relation between i and e_g' is shown in Fig. 7, together with some observed points.

TABLE II

(e_g'/e_f)	$f(e_g'/e_f)$	e_g'	i
0	0	0	0
2	4.65	18	0.024
4	16.51	36	0.084
6	32.5	54	0.166
8	53.0	72	0.270

Amplification constant calculated by equation (13) given later is $k=12.7$

$$\left(L_g=8.3 \qquad \frac{x_a}{x_g}=2.7 \qquad c=1.8\right)$$

while observed value is $k=12.8$ at $e_a=1{,}000$ and $e_g=-45$.

To draw i versus e_g characteristic curves, first assume $e_a=0$; then $e_g'=\{k/(1+k)\}e_g$ or $e_g=1.08\ e_g'$ and Table III shows the result of calculation of the characteristic for $e_a=0$, which is depicted in Fig. 8.

TABLE III

$e_a = 0$

i	e_g'	e_g
0	0	0
0.024	18	19.5
0.084	36	39
0.166	54	58
0.270	72	78

For other anode voltages, this curve is to be shifted to the negative direction by amounts as shown in Table IV.

TABLE IV

e_a	e_a/k
500	39
1,000	79
1,500	118

The resulting characteristic curves are drawn in Fig. 8 and are consistent with the observed data.

The electron current i is the sum of anode and grid currents,

$$i = i_a + i_g. \tag{10}$$

As long as the grid is negative $i_g = 0$ and $i_a = i$, but when grid becomes positive, grid current begins to flow. In ordinary tubes, and at low grid voltages where no remarkable effect of secondary emission appears, the following formula[5] is applicable for estimation of the grid current:

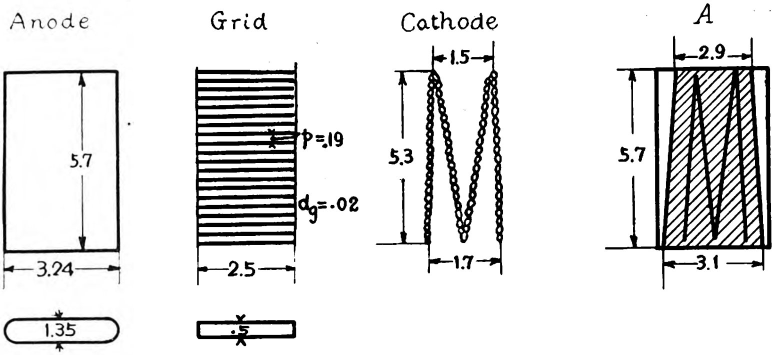

Fig. 6—Dimensions and effective anode area of type 211A tube.

$$i_g = \gamma a \sqrt{e_g/e_a} \qquad i_a = \frac{\gamma a \sqrt{e_g/e_a}}{1 + \gamma a \sqrt{e_g/e_a}}\, i \tag{11}$$

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[5] H. Lange, *Z. f. Hochfreq.*, 31, 105; April, 1928.

and hence

$$i_a = \frac{i}{1+\gamma a\sqrt{e_g/e_a}} \tag{12}$$

in which

$$\gamma = \sqrt{\log\frac{x_a}{r_c}\Big/\log\frac{x_g}{r_c}}$$

(γ_c = radius of the cathode) are usually of the order: $\gamma = 1 \sim 2 \doteqdot 1.5$ and a = ratio of the grid conductor projected area to the grid surface area.

At $e_a = 0$, the electron current is totally absorbed in the grid if the grid voltage is positive.

When the grid voltage becomes so high as to approach the anode voltage, the characteristics are severely affected by secondary emissions from the electrodes, and grid current usually increases very rapidly while anode current falls off as shown in Fig. 9.

Example 4: Calculation of grid current.

TABLE V

Tube	Observed data				Calculated grid current $i_g = 1.5\,a\sqrt{e_g/e_a}\cdot i_a$
	e_a	e_g	i_a	i_g	
102-D $a=0.156$ $k=30$	100	5.2	0.0046	0.0003	0.00025
	100	8.3	0.0074	0.0004	0.00050
	148	5.2	0.0065	0.0003	0.00029
	148	8.3	0.0100	0.0004	0.00055
UV-199 $a=0.108$ $k=6.6$	20	6	0.0023	0.0002	0.00020
	20	16	0.0040	0.0005	0.00058
	60	10	0.0038	0.0003	0.00025
	100	16	0.0042	0.0003	0.00027
UV-201A $a=0.105$ $k=7$	20	5	0.0025	0.0001	0.0002
	20	15	0.0095	0.0013	0.0013
	60	5	0.0065	0.0001	0.0003
	60	10	0.0113	0.0004	0.0007
UV-204A $a=0.16$ $k=25$	1,000	30	0.172	0.011	0.006
	1,000	60	0.274	0.019	0.009
	5,000	21	0.220	0.009	0.006
MT-9 $a=0.245$ $k=90$	1,000	80	0.305	0.027	0.031
	3,000	40	0.223	0.012	0.0094
	5,000	20	0.230	0.003	0.0033

AMPLIFICATION CONSTANT

Notations regarding the electrode dimensions are illustrated in Fig. 10.

For $r_g \ll p \ll x_g$, and for plane-electrode tubes,[6]

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[6] W. Schottkey, *Archiv f. Elektrot.*, 8, 21, 1919.

$$k=\frac{2\pi x_g}{p}\cdot\frac{\dfrac{x_a}{x_g}-1}{\log e\dfrac{p}{2\pi r_g}} \qquad (k-1)$$

and for cylindrical-electrode tubes,[7]

$$k=\frac{2\pi x_g}{p}\cdot\frac{\log\dfrac{x_a}{x_g}}{\log\dfrac{p}{2\pi r_g}}. \qquad (k-2)$$

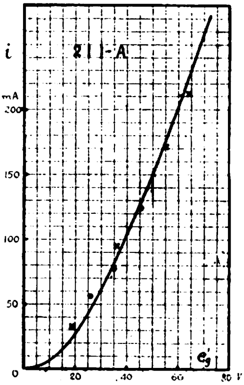


Fig. 7—Calculated characteristic of 211A.

Observed points $\begin{cases} x \text{ at } e_a = \ \ 500 \text{ v} \\ \bullet \ \ \text{"} \ \ \text{"} \ =1000 \ \text{"} \\ 0 \ \ \text{"} \ \ \text{"} \ =1500 \ \text{"} \end{cases}$

If the condition $r_g \ll p \ll x_g$ is not fulfilled, the more general formula for cylindrical-electrode tubes is to be applied,[8]

$$k=\frac{\dfrac{2\pi x_g}{p}\log\dfrac{x_a}{x_g}-\log\cosh\dfrac{2\pi r_g}{p}}{\log\cosh\dfrac{2\pi r_g}{p}-\log\sinh\dfrac{2\pi r_g}{p}}. \qquad (k-3)$$

The formulas for cylindrical electrodes were originally derived for grid wires parallel to the filament, but they are equally applicable to grids of a spiral or a system of parallel wires spanned crossways to the length of the filament.

[7] M. Abraham, *Archiv f. Elektrot.*, **8**, 42, 1919. R. W. King, *Phys. Rev.*, **15**, 256, 1920.

[8] F. B. Vogdes, *Phys. Rev.*, **24**, 683, 1924; **25**, 255, 1925.

Example 5: Calculation of amplification constant of a triode type *ABT* (made at the Laboratory), which has dimensions as shown in Fig. 11.

$$x_a = 0.20 \qquad x_g = 0.14 \qquad p = 0.12 \qquad r_g = 0.0085$$

$$\frac{2\pi x_g}{p}\ 7.3 \qquad \frac{2\pi r_g}{p} = 0.44 \qquad \log\frac{x_a}{x_g} = 0.155$$

By $(k-2)$:

$$k = 7.3 \times \frac{0.155}{0.357} = 3.2$$

By $(k-3)$:

$$k = \frac{7.3 \times 0.155 - 0.040}{0.040 + 0.345} = 2.8.$$

Observed value of $k = 2.9$.

In an actual triode in which anode and grid are plane, the cathode is not actually plane but is a filament arranged on a plane surface and accordingly does not strictly conform to the condition from which the formula $(k-1)$ has been derived. But by trying the formulas on actual cases, the writer found that the formula for plane $(k-1)$ is applicable

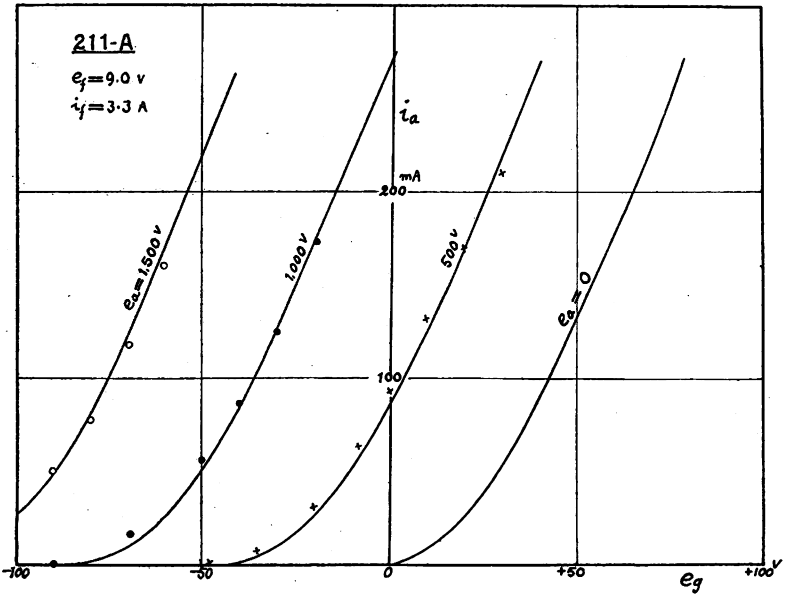


Fig. 8—Characteristics of 211A.
Calculated——
Observed at $e_a = \begin{cases} 1500 & ○ \\ 1000 & • \\ 500 & x \end{cases}$

for a tube with W-shaped filaments, while it gives somewhat larger value for tubes with V-shaped filaments. The formulas for cylinder $(k-2)$, if applied to these tubes, give lower values than observed. Comparing the two formulas, we find that they are different only in

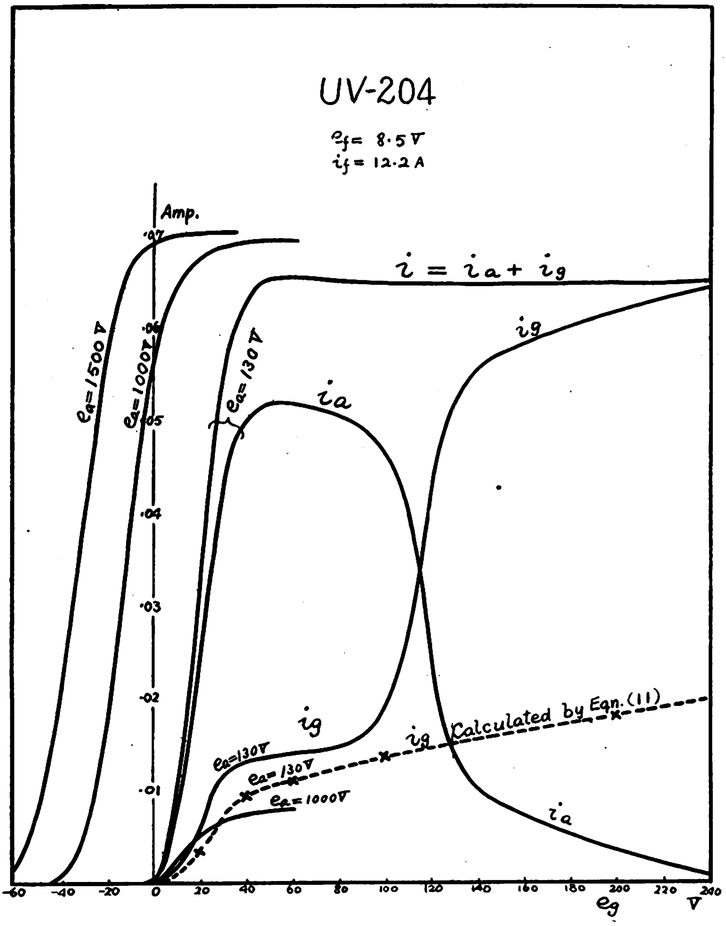


Fig. 9—Observed characteristics of UV204 showing the effect of secondary emission.

terms of x_a/x_g, and the writer suggests that for approximate calculation of k for plane-electrode tubes, the formula for cylindrical electrodes may be applied by multiplying a factor c which depends on x_a/x_g and the shape of the filament as given in Table VI.

TABLE VI

VALUES OF C

Cylindrical-electrode tubes; $C=1$
Plane-electrode tubes;
with grid of a mesh: $C=1$
with grid of parallel wires:

x_a/x_g	C W-shaped filament	C V-shaped filament
7	1.0	1.0
2	1.5	1.3
3	1.9	1.6
4	2.4	1.9
5	2.8	2.1

The formula $(k-3)$ is a more general form of $(k-2)$, and this is to be taken as a standard formula. In this formula $(\log\cosh 2\pi r_g/p)$ is of the order of 0.1 and is small compared with $(2\pi x_g/p \log x_a/x_g)$ which is usually of the order of 10 to 100, and so the formula may be reduced to a simpler form:

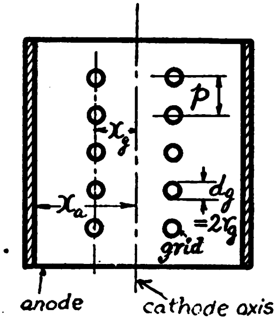


Fig. 10

$$k=\frac{c\dfrac{2\pi x_g}{p}\log\dfrac{x_a}{x_g}}{\log\coth\dfrac{2\pi r_g}{p}} \cdots \qquad (k-4)$$

Example 6: Calculation of amplification constant of a triode type UV203 (Tokyo Denki Co.) which is a plane-electrode tube with a cathode of W-shaped filament as shown in Fig. 12.

$$x_a=0.49 \qquad x_g=0.19 \qquad p=0.12 \qquad r_g=0.0085$$

$$\frac{2\pi x_g}{p}=9.95 \qquad \frac{2\pi r_g}{p}=0.445 \qquad \frac{p}{2\pi r_g}=2.25 \qquad \log\frac{x_a}{x_g}=0.412.$$

By equation $(k-1)$,

$$k=9.95\times\frac{1.60}{2.34\times 0.353}=19.6$$

and by equation $(k-2)$,

$$k=9.95\times\frac{0.412}{0.353}=11.6.$$

Equation $(k-4)$ gives

$$k=\frac{9.95\times 0.412\times 1.74}{0.467}=15.3 \quad (c=1.74)$$

and the observed value $k=15.9$.

Example 7: Calculation of amplification constant of a triode type UV201A (R. C. A.), dimensions of which are given in Fig. 13.

$$x_a=0.34 \qquad x_g=0.17 \qquad p=0.102 \qquad r_g=0.0053$$

$$\frac{2\pi x_g}{p}=10.5 \quad \frac{2\pi r_g}{p}=0.33 \quad \frac{p}{2\pi r_g}=3.03 \quad \log\frac{x_a}{x_g}=0.301$$

By equation $(k-1)$,

$$k=10.5\times\frac{1.0}{2.30\times 0.482}=9.4$$

by $(k-2)$,

$$k=10.5\times\frac{0.301}{1.022}=3.1$$

and by $(k-4)$, as the filament is V-shaped $c=1.3$ for $x_a/x_g=2.0$ and

$$k=\frac{10.5\times 0.301\times 1.30}{0.505}=8.1.$$

Observed value, $k=8.2$.

In the above formulas grid is considered to be formed of parallel wires only. But in actual cases supports are usually added and in some cases grid is constructed of wires weaved into a mesh.

The writer brought out the following deductions in order to make the formula equally applicable to these cases.

In formula $(k-4)$, the quantity $2\pi x_g/p$ means the "total active length of grid wires per unit axial length of the grid" in a cylindrical-electrode tube, because $2\pi x_g$ is the length of one turn of grid wire in circular form and $1/p$ is the number of turns per unit length. This quantity is to be denoted by L_g, thus for a grid of closely-wound spiral wire

$$L_g=\frac{2\pi x_g}{p}.$$

Example 8: Amplification constant of a triode type RE-84 (Telefunken) which has a grid of spiral wire of wide pitch of winding. (Fig. 14.) If the value of $L_g=2\pi x_g/p$ is taken, i.e. if the grid is assumed to be formed of circular rings spaced at the pitch p,

$$x_a=0.263 \quad x_g=0.125 \quad p=0.25 \quad r_g=0.025 \quad \frac{2\pi x_g}{p}=3.14$$

$$k=\frac{3.14\times 0.323}{0.25}=3.96.$$

If, on the other hand, actual length of the spiral grid wire per unit axial length of the grid as defined above is taken,

$$L_g=\sqrt{1+\left(\frac{2\pi x_g}{p}\right)^2}=3.30$$

$$k=\frac{3.30\times 0.323}{0.25}=4.4.$$

The observed value $k=4.5$ proves that the latter is more consistent.

For a grid consisting of a square mesh of very thin wires

$$L_g=\frac{4\pi x_g}{p}$$

but when grid wire is not of small diameter, effect of overlapping of the wires should be considered. At each crossing point of two wires, active length of one wire is reduced by an amount equal to the diameter of the wire, and as the number of crossing points per unit axial length of the grid is $2\pi x_g/p^2$ (Fig. 15),

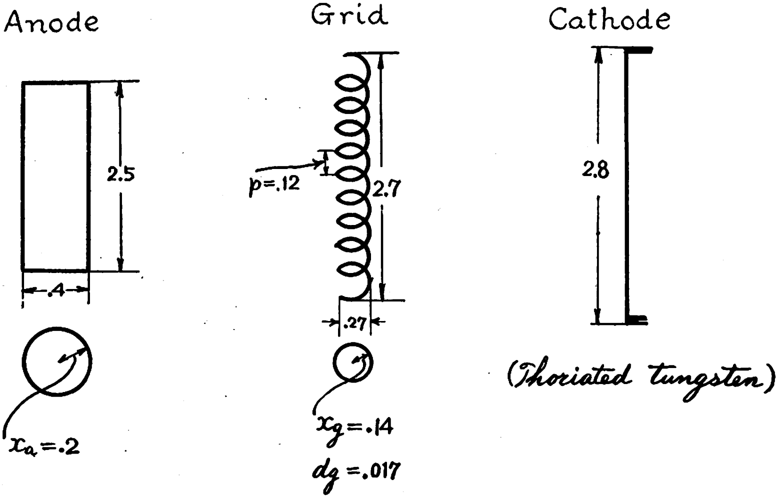


Fig. 11—Dimensions of type *ABT* tube.

$$L_g=\frac{4\pi x_g}{p}-\frac{2\pi x_g}{p^2}\cdot d_g=\frac{4\pi x_g}{p}\left(1-\frac{r_g}{p}\right).$$

Effect of grid supports should be considered in the same way and for a grid of parallel wires,

$$L_g = \frac{2\pi x_g}{p} + s - \frac{st}{p}$$

in which s is the number of grid supporting wires which are active in influencing on the space charge, and st/p is the amount of reduction of effective length of the grid wires due to the grid support covering a portion of each grid wire (Fig. 16).

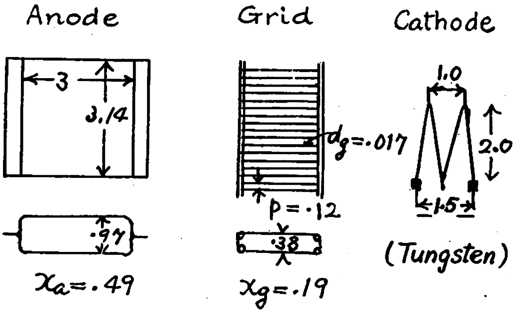

Anode
Grid
Cathode
3
3.14
.97
x_a=.49
d_g=.017
p=.12
.38
x_g=.19
1.0
2.0
1.5
(Tungsten)

Fig. 12—Dimensions of type UV203 tube.

For a grid made of a mesh, the effect of supports may not be considered, as the total length of the wires is very large, and moreover the increase of L_g due to the supports is usually canceled by the reduction of it due to the support covering a portion of the mesh.

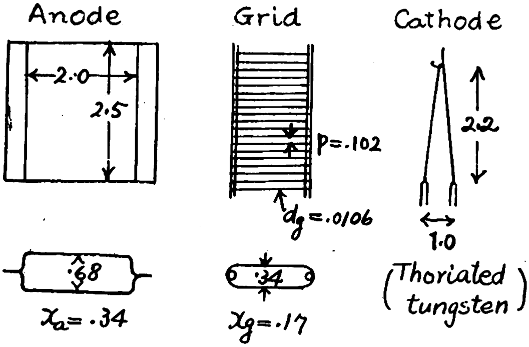

Anode
Grid
Cathode
2.0
2.5
.68
x_a=.34
p=.102
d_g=.0106
.34
x_g=.17
2.2
1.0
(Thoriated tungsten)

Fig. 13—Dimensions of type UV201A tube.

The quantity $2\pi r_g/p$ in formula $(k-4)$ signifies π-times d_g/p or "π-times the ratio of the grid conductor projected area to the grid surface area," and this ratio of areas being denoted by a,

$$\pi a = \pi \frac{\text{grid wire projected area per unit length of the grid}}{2\pi x_g}.$$

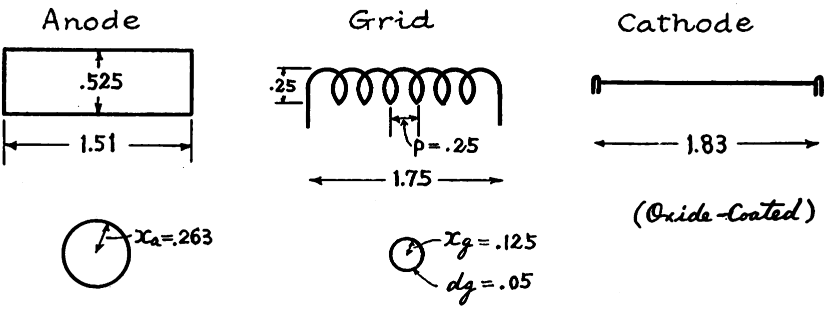

Anode
.525
1.51
x_a=.263
Grid
.25
p=.25
1.75
x_g=.125
d_g=.05
Cathode
1.83
(Oxide-Coated)

Fig. 14—Dimensions of type RE84 tube.

For a grid of a mesh or of a system of parallel wires,

$$\pi a = \frac{L_g d_g}{2x_g} = \frac{L_g r_g}{x_g}.$$

If the parallel wire grid has supports,

$$\pi a = \frac{(L_g - s)d_g + st}{2x_g}.$$

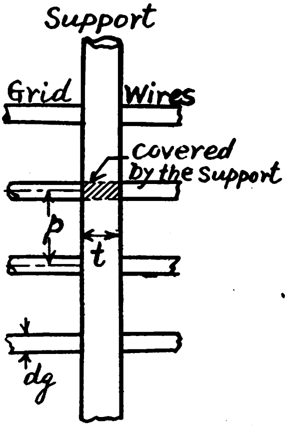

Support
Grid
Wires
covered by the Support
p
t
dg

Fig. 15

Example 9: A triode type TA08/10 (Philips' Lamp Works) has a grid supported by a thick wire. (Fig. 17.)

$$x_a = 0.64 \quad x_g = 0.24 \quad p = 0.112 \quad r_g = 0.014 \quad \log\frac{x_a}{x_g} = 0.436.$$

Without considering the support:

$$L_g=\frac{2\pi x_g}{p}=13.2 \qquad \pi a=\frac{2\pi r_g}{p}=0.79 \qquad k=31.6.$$

Considering the support: $s=1, t=0.092$

$$L_g=\frac{2\pi x_g}{p}+s\left(1-\frac{t}{p}\right)=13.4 \quad \pi a=\frac{(L_g-s)d_g+st}{2x_g}=0.93 \quad k=43.$$

Observed value: $k=46$.

Example 10: A triode type MT-4 (Marconi Co.) has a grid of a mesh as shown in Fig. 18.

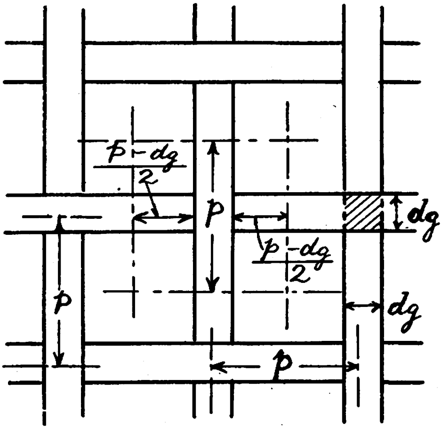


Fig. 16

$$x_a=2.17 \quad x_g=0.93 \quad p=0.13 \quad r_g=0.011 \qquad \text{og}\frac{x_a}{x_g}=0.368.$$

Without considering the overlapping of the grid wires,

$$L_g=\frac{4\pi x_g}{p}=90 \qquad \pi a=\frac{L_g r_g}{x_g}=1.06 \qquad k=\frac{90\times 0.368}{0.105}=315$$

and considering the overlapping;

$$L_g=\frac{4\pi x_g}{p}\left(1-\frac{r_g}{p}\right)=90\times 0.91=81 \quad \pi a=\frac{L_g r_g}{x_g}=0.96 \quad k=\frac{81\times 0.368}{0.129}=230.$$

Observed value: $k=210$.

The formula for the amplification constant finally attains the form;

$$k=\frac{c\,L_g \log\dfrac{x_a}{x_g}}{\log\coth \pi a}. \tag{13}$$

The estimation of L_g and πa is summarized in Table VII.

TABLE VII

Grid construction	L_g	πa
Wires parallel to the filament	n	$\frac{nd_g}{2x_g}$ or $\frac{L_g r_g}{x_g}$
Parallel wires or spiral, no support	$\frac{2\pi x_g}{p}$	$\frac{\pi d_g}{p}$ or $\frac{L_g r_g}{x_g}$
Ditto, with supports in the path of electrons	$\frac{2\pi x_g}{p}+s\left(1-\frac{t}{p}\right)$	$\frac{(L_g-s)d_g+st}{2x_g}$
Square mesh, with or without support	$\frac{4\pi x_g}{p}\left(1-\frac{r_g}{p}\right)$	$\frac{\pi d_g}{p}\left(2-\frac{d_g}{p}\right)$ or $\frac{L_g r_g}{x_g}$

These are equally applicable to tubes of either cylindrical or plane electrodes.

The amplification constant is mainly governed by grid construction and depends little on anode area and filament configuration. Main factors determining k are x_a/x_g, p/x_g, and r_g/p, and similar tubes of equal relative dimensions will give equal values of k whatever the size may be. k is affected by grid current at positive grid voltages and by the non-uniformity of grid action along the length of the cathode at very low anode currents, observed value of k being slightly reduced in both the cases.

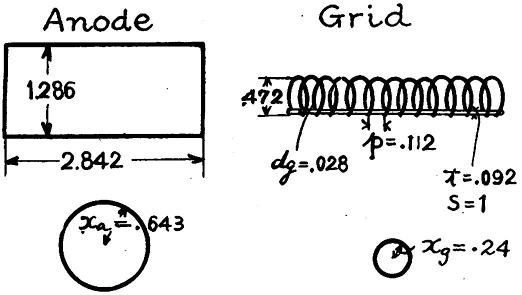


Fig. 17—Dimensions of type TA18/10 tube.

ELECTROLYTIC METHOD OF DETERMINING k

By the amplification constant k is meant that the electrostatic influence of the grid upon the cathode is k-times that of the anode. Electrostatic field in a space assumes similar distribution as stream lines of electric current that would take place if the space were filled with conducting material. The amplification constant may thus be determined by means of a model as shown in Fig. 19, in which only a single section of grid wire is taken, as this is sufficient for the purpose. By taking the ratio of electric conductance of grid-to-cathode to that of anode-to-cathode, after the bridge is balanced, the amplification constant is known as

$$k = \frac{R_a}{R_g}.$$

Table VIII shows an example obtained by the simplest structure as shown in Fig. 19, and calculated values are given for comparison.

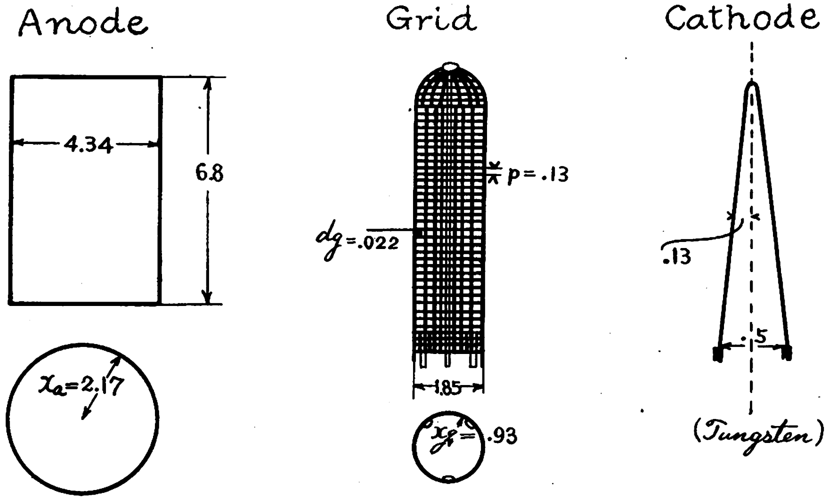


Fig. 18—Dimensions of type MT4 tube.

TABLE VIII

x_a	x_g	p	r_g	k obs.	k calc.
4.0	3.0	3.0	0.125	1.4	1.27
10.0	3.0	4.0	0.125	3.2	3.45
10.0	2.0	4.0	0.250	5.4	5.09
10.0	4.0	3.0	0.250	10.0	10.3
7.0	3.0	2.0	0.250	21.0	20.4

That the cathode configuration gives little effect on k could be evidenced by trying various forms and arrangements of the cathode conductor.

In designing a triode with very complicated structures that do not permit calculation, this method of predetermining the value of k will save much the effort of evacuation and readjustment of completed tubes.

Mutual Conductance and Anode Resistance

The mutual conductance g for given anode and grid voltages e_a and e_g can be obtained by differentiation of the characteristic equation, thus

$$g=\frac{\partial i_a}{\partial e_g}=\tfrac{2}{5}G\frac{1}{e_f}\cdot\frac{\partial}{\partial e_g}\left[e_g'^{5/2}-(e_g'-e_f)^{5/2}\right]$$

$$\fallingdotseq 1.5G\frac{k}{1+k}\left(\sqrt{e_g'}\frac{1}{4}\cdot\frac{e_f}{\sqrt{e_g'}}\right) \tag{14}$$

in which

$$e_g'=\frac{e_a+ke_g}{1+k}.$$

This is strictly applicable when grid current is not appreciable.

The mutual conductance is principally governed by the perveance G which is proportional to the anode effective area and is inversely proportional to the electrode distances.

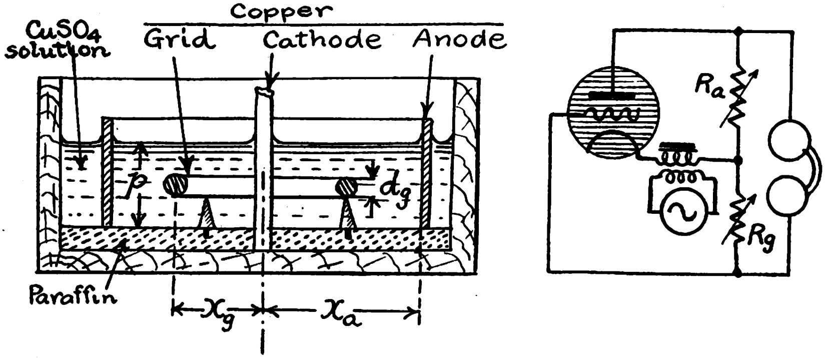


Fig. 19—Determination of the amplification constant by means of models.

The anode resistance r may also be obtained as $\partial e_a/\partial i_a$, but it is more convenient to calculate it from the known values of k and g as

$$r=\frac{k}{g} \tag{15}$$

Example 11: Calculation of mutual conductance and anode resistance of a triode type T-100 (Marconi Co.).

From the dimensions shown in Fig. 20.

$x_g=1.15$ $x_g=0.44$ $p=0.095$ $r_g=0.0075$ $A=2\pi xal=23.9$

$k_{\text{calc.}}=52.7$ $Z_a=1.15-0.08=1.07$ $Z_g=0.44-0.08=0.36$

$e_f=10$ (rating) $G=0.127\times10^{-3}$ $1.5G\frac{k}{1+k}=0.187\times10^{-3}.$

(1) To obtain g and r for $e_a=1{,}500$ and $e_g=0$,

$e_g'=27.9$ $\sqrt{e_g'}=5.28$

$g=0.187\times10^{-3}\times4.81=0.00090$ $r=\frac{52.7}{0.00090}=585{,}000.$

observed values: $g=0.00089$, $r=540{,}000$.

(2) For $e_a=2{,}000$ and $e_g=-20$,

$$e_g'=17.7 \qquad \sqrt{e_g'}=4.20$$

$$g=0.000675 \qquad r=78{,}100$$

observed values: $g=0.000675$, $r=71{,}000$.

SATURATING PART OF THE CHARACTERISTIC

Saturation current I_s depends on the material of which the cathode is made, and for a given material it depends on the dimensions and working temperature of the cathode. General characteristics of the three types of cathode, i.e. tungsten, thoriated tungsten, and oxide-coated cathode, were already published by many authorities and no particular mention is made here.

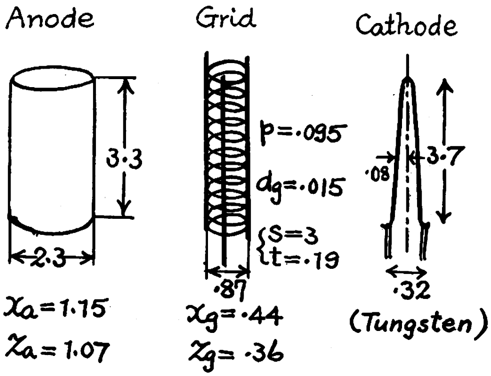


Fig. 20—Dimensions of type T100 tube.

The point at which the anode current begins to saturate occurs at an equivalent voltage e_g' such as

$$e_s=\left(\frac{I_s}{G}\right)^{2/3}.$$

If the cathode consists of a uniformly emissive surface of equal potential all over it, the characteristic would sharply bend to saturate as soon as the voltage reaches the saturating voltage e_s. But in actual cases, slow bending of the characteristic curve is invariably observed, which is caused by the non-uniformity of filament temperature as well as the voltage drop along the filament. The expression of electron current considering the latter cause only is as follows;[3] in a range of the voltage tending to saturation: $e_s<e_g'<e_s+e_f$

$$i=\frac{1}{5}G\frac{1}{e_f}[e_s^{3/2}(5e_g'-3e_s)-2(e_g'-e_f)^{5/2}].$$

This expression does not fully represent the saturating part in actual cases, and the non-uniformity of emissive power of the cathode, especially due to the cooling of the ends, is by far the important factor. Even in an indirectly-heated tube the bending takes place very slowly.

Calculation of this part of the characteristic is very difficult as the temperature distribution must be known, and moreover the emissive power of the cathode is affected by various unexpected causes. The writer finds no need of calculating this part as it is of little practical importance for the designing.

In the original paper numerical examples of computation of characteristics on about fifty types of diodes and triodes were given, being compared with their observed characteristics.

Part II: Determination of Operations of a Triode from its Static Characteristics

WORKING ANODE CURRENT

The operations of a triode as an amplifier, oscillator, modulator, etc., can be determined from its static characteristics for a given circuit arrangement, and strict and elaborated theories in this respect have been published by many authors. The following discussion is intended only for a rough estimation of the operating conditions, simplicity being aimed at with an accuracy sufficient for the designing purpose.

The static characteristic of a triode comprises two parts:

1st Part: the anode current is determined from the approximate relation,

$$i_a=Ge_g'^{3/2}=\frac{G}{(1+k)^{3/2}}(e_a+ke_g)^{3/2}=G'(e_g+ke_g)^{3/2}. \tag{15}$$

2nd Part: anode current reaches saturation, and forms the upper limit of the characteristic.

The characteristic curve is assumed to have a shape as depicted in Fig. 21 with its sharp bending point at the saturating voltage E_s. When the characteristics are simplified in such a manner, all the tubes will have the similar curves, and if we express the anode current in the unit of the saturation current I_s, and all voltages in the unit of saturating voltage E_s, a simple and universal representation of the triode characteristic is obtained.

Let E_g = grid steady voltage,
E_a = anode steady voltage,
$\mathcal{E}_g$ = grid alternating voltage in maximum value,
$\mathcal{E}_a$ = anode alternating voltage in maximum value,

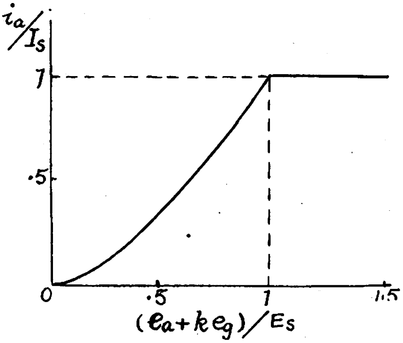


Fig. 21

positive directions being taken in the direction of direct current (Fig. 22). The load connected in the anode circuit is assumed to act as a resistance R at the operating frequency and to produce no potential drop by the d-c component of the anode current.

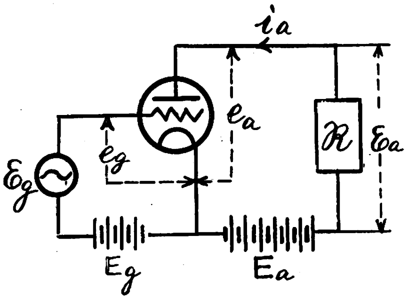


Fig. 22

In ordinary working conditions $\mathcal{E}_a$ and $\mathcal{E}_g$ are generally in opposite phase, and if the instantaneous value of the grid voltage is represented by

$$e_g = E_g + \mathcal{E}_g \cos \omega t$$

anode voltage will be

$$e_a = E_a - \mathcal{E}_a \cos \omega t$$

and for a given tube, instantaneous value of anode current will be determined by

$$e_a+ke_g=(E_a+kE_g)+(k\mathcal{E}_g-\mathcal{E}_a)\cos\omega t.$$

Thus in Fig. 23, a steady voltage E_a+kE_g determines the working point p on the characteristic and the alternating component $k\mathcal{E}_g-\mathcal{E}_a$ gives the amplitude of oscillation, the working point being the center.

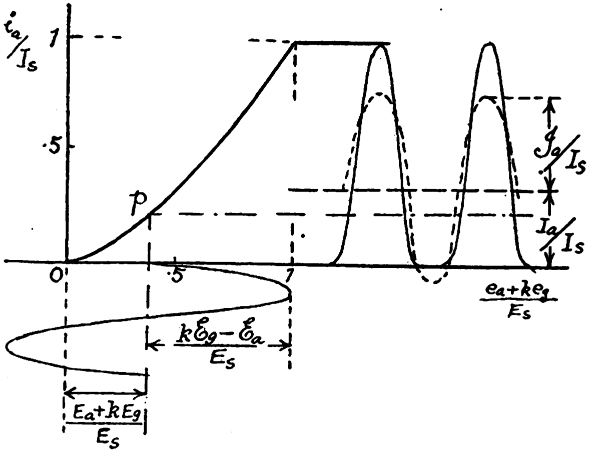


Fig. 23*

The variation of anode current due to this oscillating voltage is sinusoidal when the amplitude of voltage is very small, but becomes distorted when it is large.

Let I_a = mean value or d-c component of the anode current,
$\mathbf{I}_a$ = fundamental a-c component of the anode current, in maximum value, positive direction being taken as that of direct current.

For an infinitesimal amplitude of oscillation,

$$\mathbf{I}_a=\frac{di_a}{d(e_a+ke_g)}(k\mathcal{E}_g-\mathcal{E}_a)$$

and for a constant anode voltage,

$$\frac{di_a}{d(e_a+ke_g)}=\frac{1}{k}\cdot\frac{di_a}{de_g}=\frac{g}{k}=\frac{1}{r}$$

and

$$\mathbf{I}_a=\frac{k\mathcal{E}_g-\mathcal{E}_a}{r}$$

this gives the well-known fact that the triode circuit can be represented by an equivalent circuit as shown in Fig. 24, regarding the a-c components of working quantities.

---

* The symbol **I*a*** in the illustrations is designated by **I*a*** in the text.

For general cases, the writer obtained the wave forms of the anode current, and calculated the d-c and a-c components of the anode current by the Fourier analysis. (Fig. 23)

DYNAMIC CHARACTERISTIC DIAGRAM

The result of calculation of the d-c and a-c components of the anode current at various working voltages is summarized in a diagram shown in Fig. 25. This diagram is universally applicable to any particular

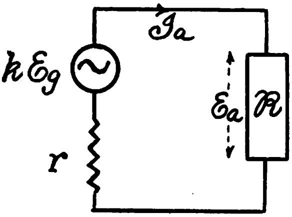


Fig. 24

triode as long as voltages are expressed in the unit of its saturating voltage E_s, and currents in the unit of the saturation current I_s. These quantities taken as units can be obtained from the static characteristics and E_s can also be calculated from the following relations:

$$E_s=\left(\frac{I_s}{G'}\right)^{2/3}. \qquad G'=\frac{G}{(1+k)^{2/3}}.$$

The dynamic characteristic diagram represents the surfaces of $\mathbf{I}_a/I_s$ and I_a/I_s as determined by the voltages $(E_a+kE_g)/E_s$ and $(k\mathcal{E}_g-\mathcal{E}_a)/E_s$, and comprises the following regions of practical importance, as designated in Fig. 26.

TABLE IX

Regions	Working point	Working range of anode current	Practical value
A	On the slope of the static characteristic	Entirely in the slope, not limited	Distortionless amplification
B	Ditto.	Limited by zero line.	Self-oscillation practicable and effectively subjected to modulation.
C	Ditto.	Limited by saturation.	Ditto, but with lower efficiency.
D	Ditto.	Limited by both zero-line and saturation.	Self-oscillation or amplification with large power.
E	At zero anode current with highly negative grid	Lower half-cycle entirely suppressed, upper peak not limited.	Power amplification of modulated waves.
F	Ditto.	Ditto, upper peak limited by saturation.	Self-oscillation or amplification of large power with high efficiency.

It should be remembered that this diagram has been derived under the following assumptions:

(a) static characteristic has sharp bending toward saturation,

(b) anode and grid voltages vary in sine wave form and just in opposite phase,

(c) grid current is not appreciable,

(d) load connected in the anode circuit does not produce appreciable voltage drop by direct current as well as by harmonic waves.

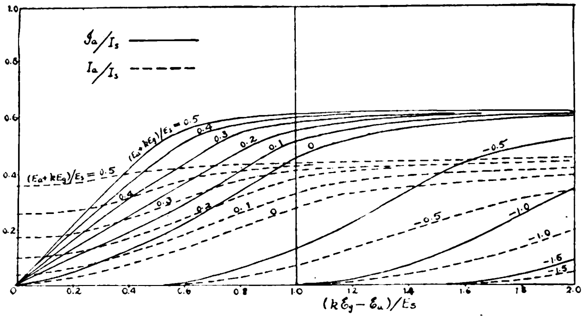


Fig. 25—Dynamic characteristics.

Thus the diagram is not strictly applicable when there is considerable grid current flowing, which takes place when the maximum grid voltage $E_g+\mathcal{E}_g$ becomes comparable with the minimum anode voltage $E_a-\mathcal{E}_a$. In such a case the anode alternating current attains a drooping characteristic at high excitation as shown in Fig. 27.

EXPERIMENTAL VERIFICATION

The dynamic characteristic diagram was obtained by experiment in the following way. A triode was used as a radio-frequency power amplifier in a circuit shown in Fig. 28. The static characteristics of the tube are shown in Fig. 29, from which the necessary quantities for the units were calculated as follows:

$$I_s=0.071 \qquad E_s=1.220 \qquad k=25.$$

During the experiment, under constant filament current, E_a and E_g were varied and at each step of variation, I_0 and I_a were observed with variation of I' and C_g. Calculations were then made as follows,

$$\mathcal{E}_a=\frac{\mathrm{I}_0}{\omega C} \qquad \mathcal{R}=\frac{L}{CR}=\frac{1}{\omega^2C^2R} \qquad \mathrm{I}_a=\frac{\mathcal{E}_a}{\mathcal{R}} \quad \mathcal{E}_a=\frac{\mathrm{I}'}{\omega C_g}\cdot$$

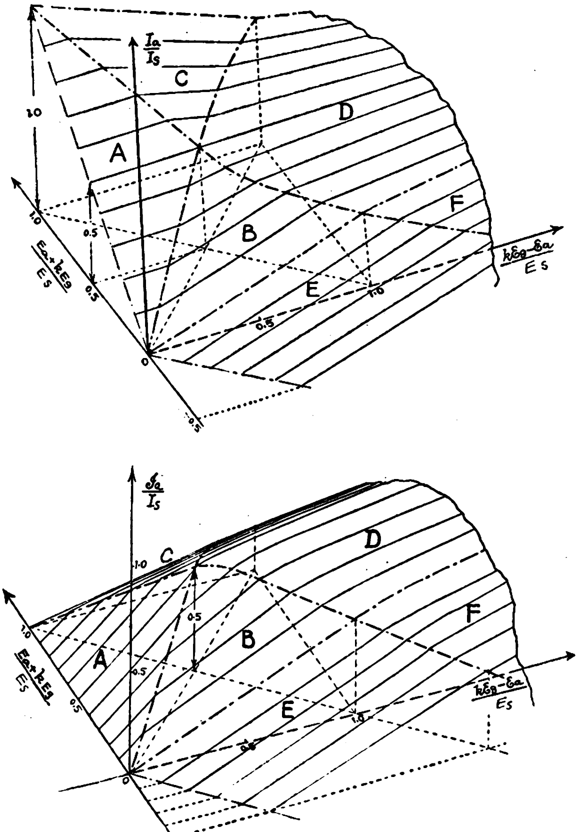
In the actual case, $\omega=3\times10^6(\lambda=628\text{ m})$ and high-frequency resistance of the oscillatory circuit was 3 ohms so that $R=3+R'$. Hence

$(k\mathcal{E}_g - \mathcal{E}_a)/E_s$ could be calculated for each value of $(E_a + kE_g)/E_s$, and I_a/I_s and I_a/I_s were plotted against $(k\mathcal{E}_g - \mathcal{E}_a)/E_s$ to obtain the dynamic characteristic curves.

The resulting curves are shown in Fig. 30, which are in conformity with the theoretical diagram in Fig. 25 in the main.

APPLICATION OF THE DIAGRAM

In Fig. 31, the dynamic characteristic diagram is shown with its left side extended to another diagram in which the ordinate gives



Ia/Is
C
D
A
B
E
F
1.0
0.5
1.0
(Ea+kEg)/Es
0.5
0
1.0
0.5
(kEg−Ea)/Es
0.5
Ia/Is
1.0
C
D
0.5
A
B
F
0.5
1.0
(Ea+kEg)/Es
0.5
E
1.0
0.5
(kEg−Ea)/Es
0
0.5

Fig. 26—Dynamic characteristic surfaces.

currents in the unit of I_s, and the abscissa gives voltages in the unit of E_s. In this part of the diagram a group of straight lines starting at the origin represent the external resistance R in the unit of the mean anode internal resistance $\bar{r}$ which is derived as $\bar{r} = E_s/I_s$, and hyperbolas give the contour lines of power, unit of which is to be taken as E_sI_s.

To explain the use of the diagram, assume that a triode having saturation current I_s, amplification constant k, and saturating voltage E_s, is to be used as an amplifier with an anode circuit impedance R, and its operating conditions are to be studied. First calculate the unit quantities $\bar{r} = E_s/I_s$ and E_sI_s from the given E_s and I_s. All the currents, voltages, resistances and powers are to be expressed in the units of I_s, E_s, $\bar{r}$ and E_sI_s, respectively. In the following explanations quantities expressed in this manner are taken instead of actual values.

Let the anode voltage be E_a and grid voltage be E_g. Calculate $E_a + kE_g$ and the corresponding dynamic characteristic curves of $\mathbf{I}_a/I_s$ and I_a/I_s are found on the diagram such as the lines *oeg* and *def* in Fig. 32. Draw a line *oh* to represent the given external impedance. For a given grid exciting voltage $\mathcal{E}_g$, calculate $k\mathcal{E}_g$ and take this quantity on the right abscissa such as the point i; from i a straight line is drawn parallel to the resistance line *oh* and let this line intersect with the dynamic line *oeg* at p; then this point p gives the working point on the dynamic characteristic. The reason is as follows: the ordinate of p represents $\mathbf{I}_a$ and let a level line through p intersect with the *oh* line at j and with the ordinate axis at k, then $\overline{ok} = \mathbf{I}_a$ and $\tan \angle jok = R$. Therefore $\overline{jk} = \mathbf{I}_aR = \mathcal{E}_a$, but $\overline{pj} = \overline{oi} = k\mathcal{E}_g$ and thus the abscissa of the point p is $\overline{pk} = \overline{pj} - \overline{jk} = k\mathcal{E}_g - \mathcal{E}_a$ and is in conformity with the dynamic characteristic curves which have been expressed by the abscissa of $k\mathcal{E}_g - \mathcal{E}_a$.

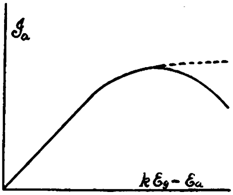


Fig. 27

The area of the triangle *ojk* gives $\frac{1}{2}\mathcal{E}_a\mathbf{I}_a$ or $\frac{1}{2}\mathbf{I}_a{}^2R$ and represents the power output, which can be read by means of the power contour hyperbolas numbered at the top. The steady component I_a is obtained by drawing a line vertically from point p to intersect with the I_a/I_s dynamic line at a point q. Take $\overline{oa}$ equal to E_a. A level line through q intersects with the vertical line through a at l, and with the ordinate axis at m. Then $\overline{om} = I_a$, and the rectangular area *omla* is E_aI_a and represents the anode power input to the tube which can be read from the position of the point l on the contour hyperbolas numbered at the left end. The efficiency of power conversion can be calculated from the two power values as $\eta = \frac{1}{2}\mathbf{I}_a{}^2 R / E_a I_a$.

When the triode is used as a self-excited oscillator,[9] the grid voltage $\mathcal{E}_g$ is in definite relation to the anode voltage $\mathcal{E}_a$ as determined from the circuit conditions. In an example shown in Fig. 33, $\mathbf{I}_a$ is determined from $\mathcal{E}_g$ by a dynamic line *ops*, while $\mathbf{I}_a$ and external impedance $R \doteq L/CR$ determine $\mathcal{E}_a = \mathbf{I}_a(L/CR)$ and this induces an e.m.f. $\mathcal{E}_g = \mathcal{E}_a(M/L)$ in the grid circuit. Hence $\mathbf{I}_a/\mathcal{E}_g = CR/M$ is a straight line, the inclination of which is $\tan^{-1} CR/M$ as shown by a line *ot* in Fig. 33. Thus the dynamic characteristic curve *ops* gives the relation in which $\mathbf{I}_a$ is determined by $\mathcal{E}_g$ from the tube characteristics, while the excitation line *ot* gives the relation in which $\mathcal{E}_g$ is determined by $\mathbf{I}_a$ from circuit conditions. Then the point of intersection *p* shall be the working point of the self-oscillator. If the coupling becomes loose, the excitation line is shifted to left and finally will not intersect with the dynamic line, and in this case oscillation cannot be produced.

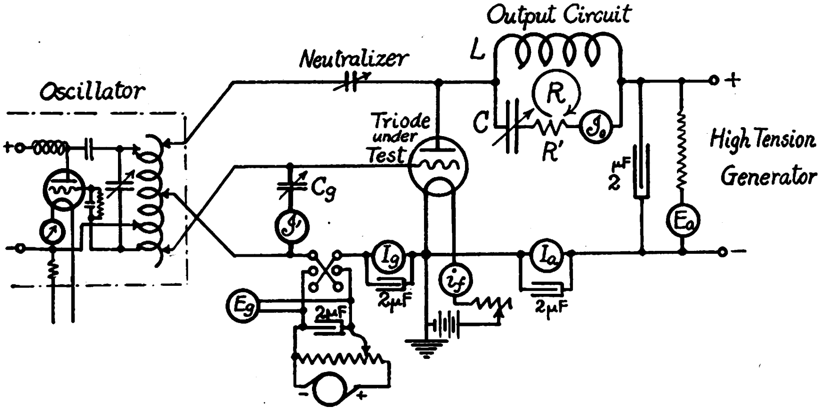


Fig. 28

The regeneration can be explained in the same diagram.[9,10] When regeneration is applied, anode circuit is coupled to grid circuit but not so close as to produce oscillation, such as a line *ot'* in Fig. 33. If the regenerative connection is not made, for a certain grid voltage such as $\overline{ou}$, the anode current is $\overline{uv}$. But on regenerative condition, a line is to be drawn from the point *u* parallel to the line *ot'* and the intersecting point *w* on the characteristic gives the working point. The reason is that the level line from *w* intersecting with *ot'* at *r* and with the ordinate axis at z represents the grid excitation which is the sum of the originally impressed voltage $\overline{ou}=\overline{rw}$ and the voltage induced by the anode current $\overline{rz}$. The ratio of regenerative amplification to non-regenerative one may be expressed by $\overline{oz}/\overline{uv}$.

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[9] H. G. Möller, "Elektronenröhren und ihre technischen Anwendungen," 1920.

[10] E. H. Lange, *Phil. Mag.*, 750; October, 1925.

Now these ideas can be applied to the dynamic characteristic diagram as shown in Fig. 34. Comparing the figure with Fig. 33, resistance line *oh* corresponds to the ordinate axis of Fig. 33, and the abscissa representing $k\mathcal{E}_g$ corresponds to that of Fig. 33. The excitation line *ot* in Fig. 33 is obtained in Fig. 34 by taking the quantity $k\mathcal{E}_g/\mathbf{I}_a$ from the point h to the right on the level line as a point t, and by connecting this point to the origin; in other words $\{(k\mathcal{E}_g/\mathbf{I}_a)-R\}/\bar{r}$ shall be taken on right abscissa as $\overline{zt}$ in order to obtain the excitation line *ot*.

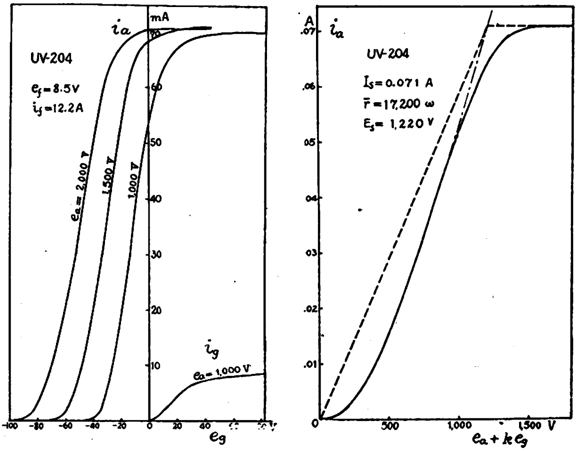

Fig. 29—Characteristics of type UV204 tube.

The point of intersection of the excitation line with the dynamic characteristic curve gives the working point p and all the other quantities can be derived in the same way as explained before. In Fig. 31 $(k\mathcal{E}_g/\mathbf{I}_a)_c$ means that the quantity is to be derived from the circuit conditions.

The modulation of a self-oscillating tube consists in altering the anode or grid steady voltage with audio frequency, hence producing the corresponding variations in the amplitude of oscillation. This means the variation of E_a+kE_g, other things being unvaried. Thus, for example, if anode voltage E_a is varied in sinusoidal wave form as shown in Fig. 35, E_a+kE_g will vary accordingly and the dynamic characteristic curve is shifted up and down. The excitation line *ot* remaining unvaried, the working point p must travel on the *ot* line and the oscillating current varies its amplitude in accordance with the variation of the anode voltage.

Applications of the diagram have been further illustrated by numerical examples in the original paper. The diagram gives a bird's-eye view of various working conditions.

Part III: The Design of Triodes

WORKING POINTS ON THE DYNAMIC CHARACTERISTICS

The design of a triode to satisfy specified conditions can be accomplished in the following way, which shows a typical case where the use of the triode is indicated and its power output is given.

Optimum working points can always be found on the dynamic characteristic diagram for any particular use of a triode. These optimum conditions which the writer conceives to be appropriate are given below.

TABLE X

Use of Triode	$(E_a+kE_g)/E_s$	I_a/I_s	R/r
Low power amplifier	0.3—0.5	0.2—0.3	1.0
Distortionless power amplifier or modulator	0.3—0.5	0.2—0.4	2.0
Oscillator or power amplifier for radio-telegraphy	—1.0	0.55	1.0
Self-oscillator for radio-telephony	0.3	0.35	1.2
Power amplifier for radio-telephony, amplifying modulated wave	0—0.2	0.25	1.0
Ditto, amplifying continuous wave and subjected to modulation	0	0.3	1.5

If the above three quantities are decided, the following may be directly obtained from the diagram:

$$\frac{P}{E_sI_s}, \quad \frac{k\mathcal{E}_g}{E_s}, \quad \frac{\mathcal{E}_a}{E_s}, \quad \frac{I_a}{I_s}.$$

Moreover E_a/E_s can be roughly estimated according to the use of the tube and in relation to the value of $\mathcal{E}_a/E_s$, and from this the approximate values of E_aI_a/E_sI_s and η can be found.

HIGH-IMPEDANCE OR LOW-IMPEDANCE TUBE

For a given power output P, $\alpha=P/E_sI_s$ having been known, E_sI_s can be obtained and next comes an essential problem of splitting E_sI_s into the saturating voltage E_s and the saturation current I_s.

If one of any other operating conditions, such as anode voltage E_a

or anode circuit impedance R, is given at the same time, the problem is directly solved.

If this is not the case, especially in the design of standard types of tubes, most precise considerations must be taken for this from technical and economical points of view. This is the problem of selecting the type of the tube between a high-voltage low-emission type, and a low-voltage high-emission type. The general considerations on the two alternatives, high-impedance tube or low-impedance tube in other words, are as follows for a given power output:

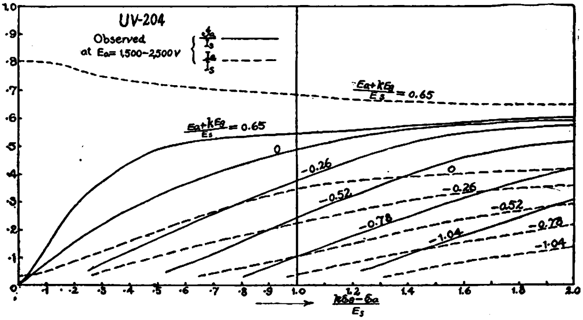

Fig. 30—Dynamic characteristics of UV204 tube obtained by experiment at radio frequency.

TABLE XI

	High impedance tube	Low impedance tube	Remarks
Anode voltage	high	low	← Anode voltage must be within a practicable limit, from standpoint of insulations of the tube and other apparatus.
Saturation current	low	high	
Cathode power consumption	little	large	← Lower filament power is preferable as overall operating efficiency becomes higher and anode radiating area can also be reduced.
Circuit impedance	high	low	
Amplification constant	high	low	

Anode power conversion efficiency is not very different when it is considered on the diagram as it is determined from the working points, but in a higher voltage tube, anode voltage is more completely utilized and net efficiency will be a little higher than lower voltage tube.

This problem involves not only the economy of the tube itself but also that of apparatus with which the tube is to be used, such as output circuits, anode source of power, amplifiers preceding the tube under consideration, etc. The practicable limits in the values of E_a, R, cathode heating power, and relative amount of cathode heating power to anode loss, should also be taken into consideration.

Regarding the oscillatory power output of a triode, the following consideration is sometimes necessary. Output P, somewhat larger than

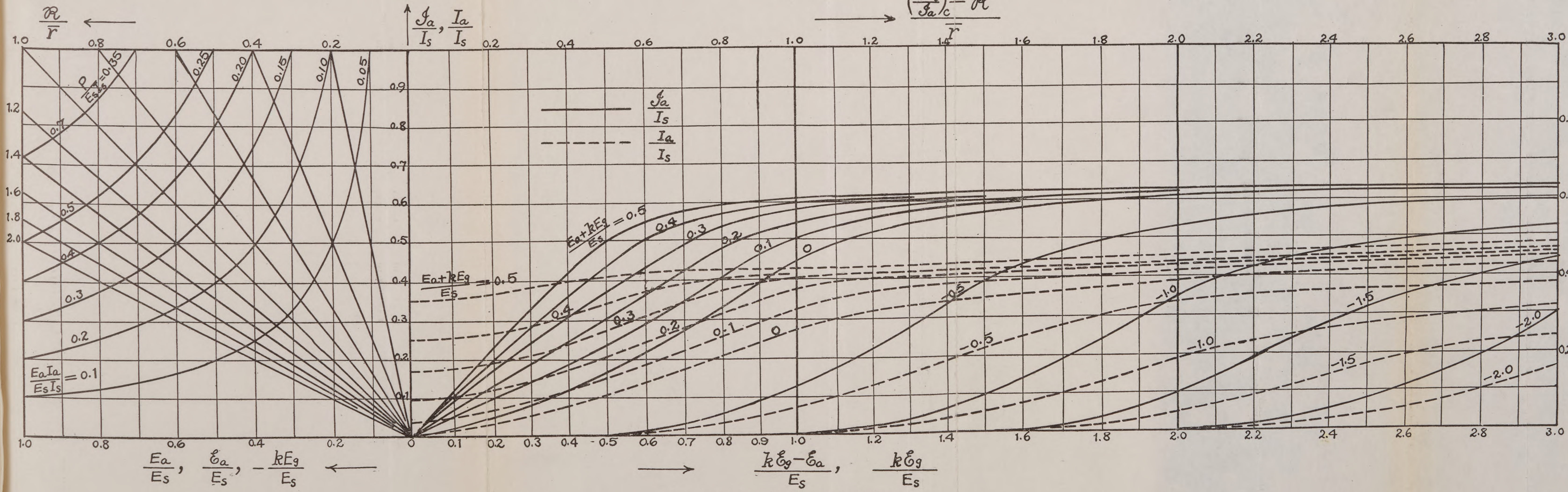

Fig. 31—Dynamic characteristic diagram.

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the required value, has to be taken as the designing basis in order to take into account effect of the grid current, which was not done in the derivation of the dynamic characteristic diagram. The grid current affects the power output in two ways; first it reduces the anode current and hence the output, and secondly it absorbs some portion of the output in case of self-oscillators. An increase of 10 or 20 per cent of the required value will generally be sufficient, according as the tube is used as a power amplifier or a self-oscillator. More rigorous estimation of grid excitation loss will be given later.

AMPLIFICATION CONSTANT

From equation (15),

$$I_s = \frac{G}{(1+k)^{3/2}} E_s^{3/2} \quad \text{or} \quad (1+k)^3 = G^2 E_s^3 / I_s^2 = G^2 \bar{r}^3 I_s \tag{16}$$

and

$$P = \alpha E_s I_s$$

hence

$$(1+k)^3 = \left(\frac{G\alpha}{P} \right)^2 E_s^5. \tag{17}$$

G is not very much different with the type of tubes and is of the order as shown in the following table. This is because G is determined by the ratio of anode area to distances between electrodes, and in practical construction of the tube there are limits in the attainable anode area for a given length of the filament and for a given span of the filament the distances between electrodes must not be brought too close together in order to prevent an accidental short-circuit of the electrodes during operation.

TABLE XII

Tubes		No. of tubes examined	G	
Anode voltage	Electrodes		Range	Average
6,000–10,000	cylinder	14	$0.08-0.4\times10^{-3}$	0.15×10^{-3}
200– 5,000	cylinder	8	$0.08-0.4$	0.17
"	plane	15	$0.15-0.9$	0.38
below 100	cylinder	10	$0.12-0.5$	0.18
"	plane	5	$0.28-0.5$	0.33

Total average: G for plane electrodes, 0.35×10^{-3}
G for cylindrical electrodes, 0.16×10^{-3}.

α and P having been known, and G being assumed, k is in definite relation to E_s as given by equation (17), and once E_s is determined k is directly obtainable.

In case of oscillators, self or separately excited, power lost in the excitation depends on the value of k. In a self-excited oscillator, this power is converted from the output of the tube and too much exciting power results in the lowering of the net output and overloading of the grid. In a separately-excited oscillator or power amplifier, the grid loss is to be supplied from the preceding tube. In these cases the value of k may be estimated as follows; for the most favorable operation delivering maximum power at a given anode voltage, the minimum anode voltage $E_a - \mathcal{E}_a$ must be nearly equal to the maximum grid voltage $E_g + \mathcal{E}_g$,[11] E_g being negative, and this requires that

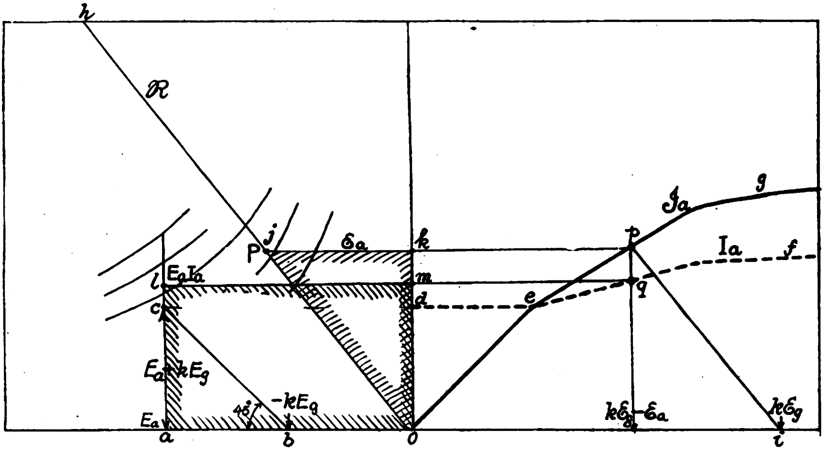

Fig. 32

$$E_a - \mathcal{E}_a = E_g + \mathcal{E}_g$$

or

$$\frac{E_a}{E_s} - \frac{\mathcal{E}_a}{E_s} = \left(\frac{kE_g}{E_s} + \frac{k\mathcal{E}_g}{E_s}\right)\frac{1}{k}$$

hence

$$k = \frac{\left(\frac{k\mathcal{E}_g}{E_s}\right) + \left(\frac{E_a + kE_g}{E_s}\right) - \left(\frac{E_a}{E_s}\right)}{\left(\frac{E_a}{E_s}\right) - \left(\frac{\mathcal{E}_a}{E_s}\right)} \tag{18}$$

in which all quantities except E_a/E_s have accurately been known. Thus if k is given, E_a/E_s can be accurately determined and E_aI_a/E_sI_s and η are accordingly known.

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[11] D. C. Prince, Proc. I. R. E., 11, 280; June, 1923.

For a rough estimation of the grid excitation loss in the above working condition, the grid current may be assumed to have a triangular wave form as shown in Fig. 36, and its peak value to reach $\frac{1}{2}I_s$. Of course this condition may not take place in actual case, but this will not be far from it. The results of calculations are as follows:

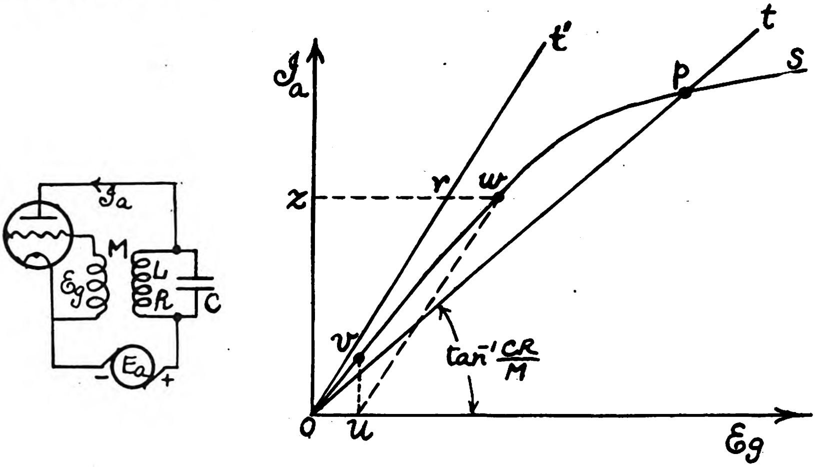

Fig. 33

Grid current d-c component, $I_g=\mu I_s$.

Grid leak loss (or loss in charging the grid biasing battery), $W_l=-\mu E_g I_s$.

Grid excitation loss, $W_e=\nu \mathcal{E}_g I_s$.

Grid loss inside the tube or grid heating power, $W_g=(\nu\mathcal{E}_g+\mu E_g)I_s$.

Grid leak resistance required, $R_g=-E_g/I_g=-E_g/\mu I_s$.

in these relations E_g is a negative quantity, and μ and ν are the functions of $-E_g/\mathcal{E}_g$ or $-(kE_g/E_s)/(k\mathcal{E}_g/E_s)$ as given in Fig. 37.

If the grid loss inside the tube be allowed up to $1/b$ of the anode loss, $1/b\,(E_aI_a-P)\geqq(\nu\mathcal{E}_g+\mu E_g)I_s$ or

$$\frac{1}{b}\left(\frac{E_aI_a}{E_sI_s}-\frac{P}{E_sI_s}\right)\geqq\left(\nu\frac{k\mathcal{E}_g}{E_s}+\mu\frac{kE_g}{E_s}\right)\frac{1}{k}$$

hence

$$k\geqq b\frac{\nu\left(\frac{k\mathcal{E}_g}{E_s}\right)+\mu\left(\frac{kE_g}{E_s}\right)}{\left(\frac{E_aI_a}{E_sI_s}\right)-\left(\frac{P}{E_sI_s}\right)}. \tag{19}$$

If, on the other hand, the grid excitation loss is to be limited within $1/d$ of the power output,

$$\frac{P}{d} \geqq \nu \mathcal{E}_g I_s \quad \text{or} \quad k \geqq d\nu\left(\frac{k\mathcal{E}_g}{E_s}\right)\Big/\left(\frac{P}{E_s I_s}\right). \tag{20}$$

All the quantities in the parentheses have been known and the lower limit of k can be determined from these relations. The exact value of E_a/E_s must be derived by equation (18), after the value of k is determined.

It is advisable to calculate the required quantities for several values of k or E_s, and to select among them a most preferable one.

DIMENSIONS OF THE ELECTRODES

The main factor determining the anode construction is the anode loss W_a which may be taken as $(E_aI_a - P)$. But in the case of amplifiers or modulators for radiotelephony, the dead anode loss corresponding to the working point on the static characteristics, i.e., $GE_a\{(E_a + kE_g)/(1+k)\}^{3/2}$ must be taken as the anode loss, as this amount of loss actually takes place when speech is not being transmitted.

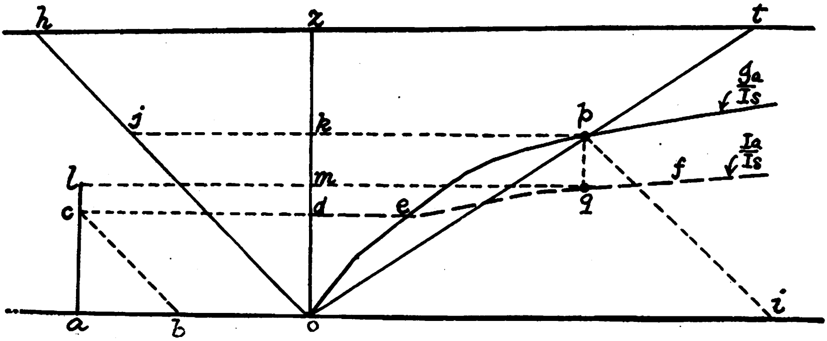

Fig. 34

The allowable limit of anode loss is dependent on the material of the anode and its cooling condition. The following data giving the maximum limit of the loss continuously applicable to the anode may be taken in the ordinary construction.

TABLE XIII

Materials	Allowable loss (watt/cm²)
Nickel	$w = 3$
Molybdenum	5
Tungsten	8
Copper, water-cooled	20

The filament heating power W_f must also be considered in estimating the anode cooling area, as some portion of it will have to dissipate through the anode. Suppose q part of the filament power is radiated through the anode, then

$$A \geqq \frac{W_a + qW_f}{w} \tag{21}$$

q depends on the degree at which the filament is hidden in the anode, and in ordinary cases in which the anode is open on both ends, q may be taken as about 0.5. The writer deems it reasonable to take the effective area A explained in Part I, instead of the actual area, in the above expression.

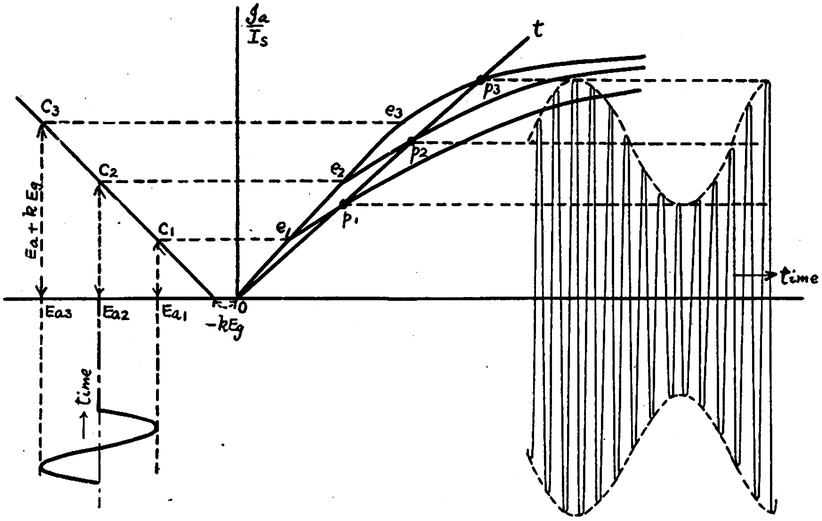

Fig. 35

In high-voltage tubes grid voltage also becomes high with respect to the cathode, and appreciable distances must be kept between grid and cathode in order to avoid severe electrostatic attraction. On the other hand, in low-voltage tubes, x_0 can be reduced, and if the cathode is made of W-shaped filament arranged in a plane, A may be greatly increased without difficulty in construction. Thus for anode voltages less than say 4,000 v, plane anode is preferable, while for higher voltages cylindrical anode is advantageously applied.

The order of x_a in the existing tubes is as follows:

TABLE XIV

E_a below 100 V	X_a =0.2–0.5
E_a 100–1,000	0.3–1.0
E_a 1,000–10,000	0.6–3.5

The actual anode surface area should be made as effective as possible in order to reduce the mass of the anode material and hence to minimize the absorbed gas evolving in the evacuation process, as well as from an economical point of view. The plate should be as thin as mechanically allowable, but not so thin as to avoid free conduction heat, otherwise local overheating or melting may occur at accidental overload of the anode.

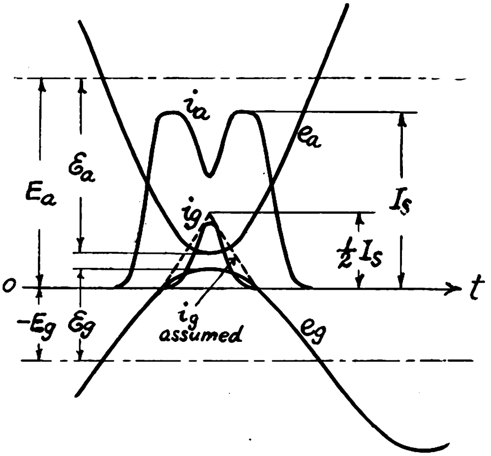

Fig. 36

The anode area A and anode-to-cathode distance x_a or z_a must be in conformity with each other to satisfy the expression,

$$G = 2.33 = 10^{-6}\frac{A}{z_a z_g}\cdot$$

The grid construction is determined from the amplification constant. When x_a is given and other dimensions p and d_g remain unvaried, k is maximum at $x_g \doteq 0.4\, x_a$, and it is advisable to take this in order to reduce the mass of the grid as well as to minimize the effect of non-uniformity of x_g on the value of k. Thus

$$\frac{k}{\log\dfrac{x_a}{x_g}} = \frac{k}{0.4} = \frac{cL_g}{\log\coth\pi a}\cdot$$

The table in Part I is to be referred to for the calculation, and ranges of actual values of L_g and πa in existing tubes are shown in Fig. 39.

L_g and πa having been determined, the next problem is to determine

p and r_g. This means a decision as to whether to use thick wires at broad pitch or to use thin wires closely arranged.

It is preferable to take πa small in order to reduce the mass of the grid and to facilitate evacuation. This is effected by thin grid wires at comparatively narrow pitch, but in extreme cases mechanical rigidity of the grid will be lost and deformation or break down will occur during operation by the heating of the grid. Fig. 38 shows actual values of p and r_g in existing tubes, from which the general tendency can be observed. Materials best suited for grid are molybdenum and nickel, but the latter has a remarkable effect of secondary emission and often causes the "blocking" of oscillators.

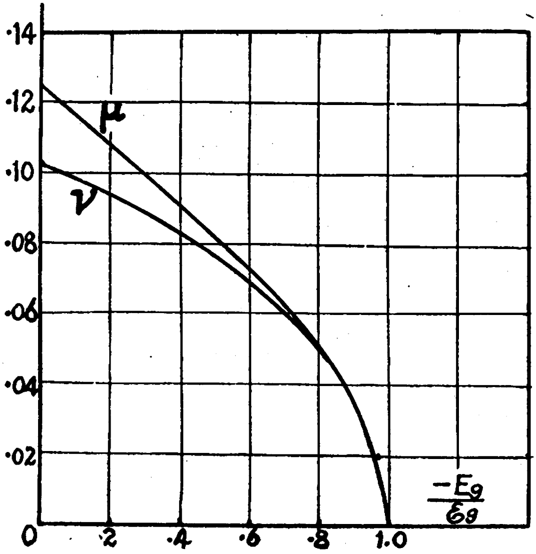

Fig. 37

DESIGN OF THE CATHODE

Design of a tungsten cathode to give the required emission is fully described in the original paper, applying a chart worked out by N. Kato[12] based on the data given by Forsythe and Worthing.[13] There have been presented various other methods on the design of cathode, and no explanation is made here.

For a close approximation of the reasonable life of a tube, which in ideal case is the life of the cathode, an economical condition must be dealt with in such a way that the total running cost of a tube is a

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[12] N. Kato, "Graphs for the Design of Bright-emitting Tungsten Filament," Circulars of the Electrotechnical Laboratory, No. 50; February, 1928. (Written in Japanese.)

[13] W. E. Forsythe and A. G. Worthing, *Astrophysical J.*, 61, 146, 1925.

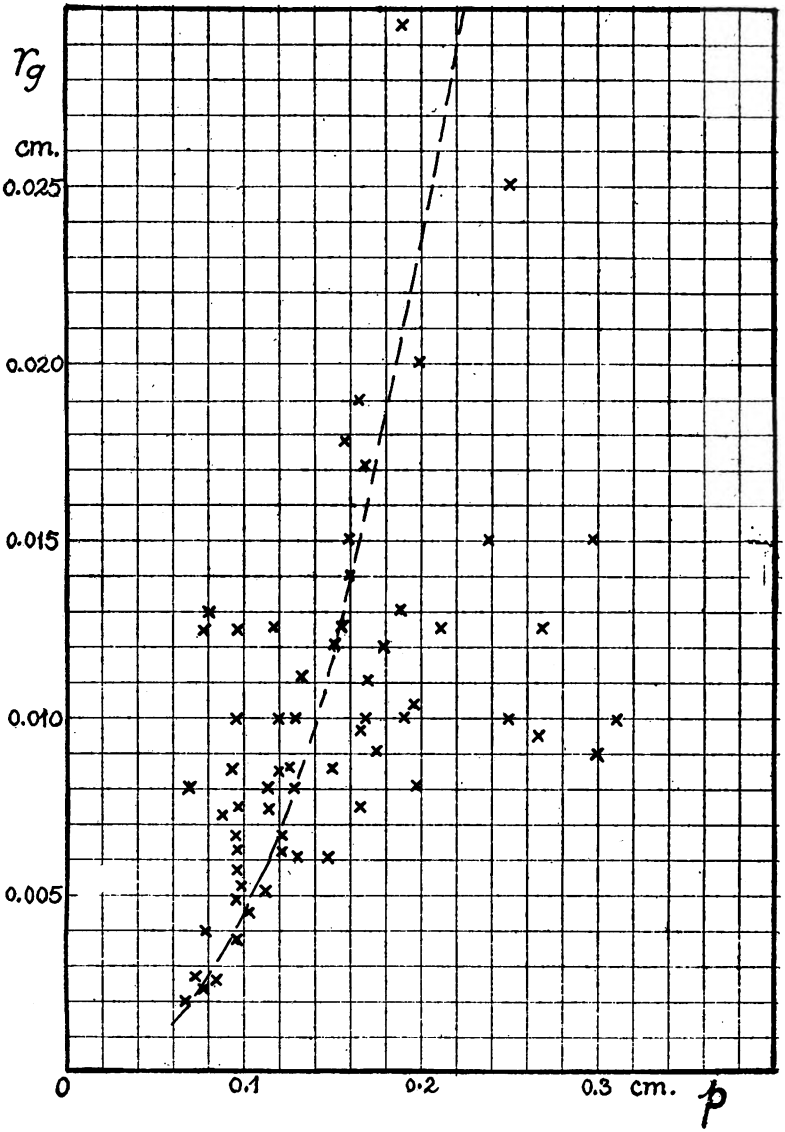

Fig. 38

minimum for a given power output. This condition is given by the following relations: total cost per watt output per hour,

$$T=\frac{Ct}{PL}+\left(\frac{1}{\eta}=\frac{W_f}{p}\right)C_g$$

in which C_t = cost of the tube, and C_p = cost of electric power per watt-hour. On the other hand, for the filament of a certain diameter the life L depends on its working temperature, which in turn is determined by the filament efficiency I_s/W_f at which it is operated, thus

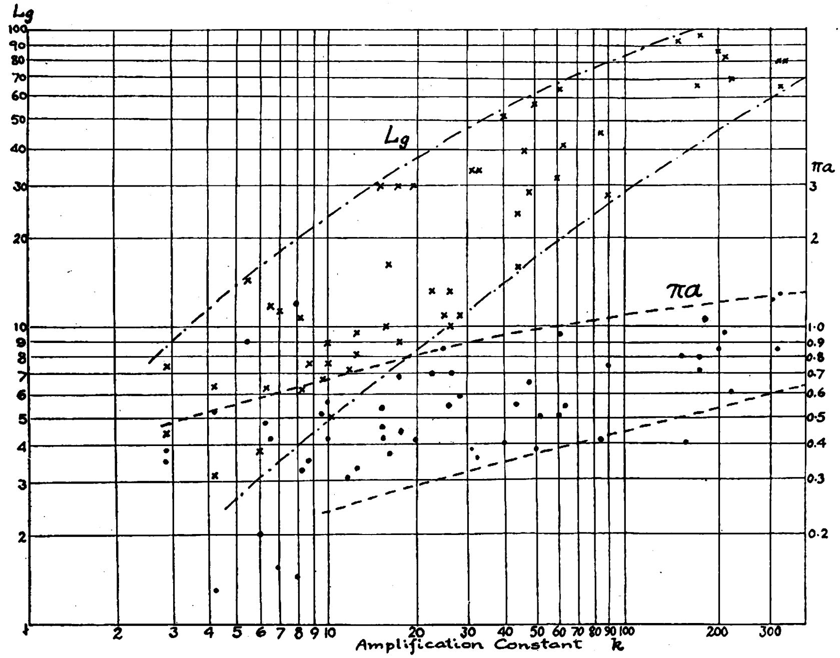

Fig. 39—Ranges of $Lg(x)$ and $\pi a(\cdot)$ in existing tubes.

$$L=\phi\left(\frac{W_f}{I_s}\right)^{\sigma}.$$

Minimum of the operating cost T occurs at a life

$$L_{\text{opt.}}=\phi^{1/1+\sigma}\left(\frac{\sigma C_t}{C_g I_s}\right)^{\sigma/1+\sigma}.$$

For tungsten filament, the two constants σ and ϕ are of the following orders,[14]

$$\sigma\fallingdotseq 2.6\quad \begin{array}{c|ccc} d_f & 0.005 & 0.020 & 0.050 \\ \hline \phi & 0.00034 & 0.0026 & 0.033 \end{array}.$$

[14] H. Simon, *Telefunken Z.*, Fig. 44 on p. 39; October, 1927.

It is preferable to design the filament to give a slightly higher value of emission than desired in order to give allowance for the discrepancies which might take place, as the emission is quite a variable quantity.

For tungsten filaments it is sometimes necessary to take a slightly longer life than calculated, as a portion of the designed life will be lost during the evacuating process by evaporation.

CONCLUSION

In the above discussions the fundamental principles involved in the writer's system of designing is explained. Many of the factors that have to be dealt with from the practical point of view in the manufacturing are lacking, and further development is necessary in this respect, but it is hoped that this paper will be of some use to manufacturers.

The writer wishes to acknowledge his indebtedness to E. Yokoyama, under whose direction the present work has been carried out, and also to N. Kato, I. Miura, and S. Maeda for their assistance.

List of Symbols

A	effective anode area (cm^2)
a	ratio of grid conductor area to grid surface area
c	conversion factor of k for plane-electrode tubes
d_f	filament diameter (cm)
d_g	grid wire diameter (cm)
e_a	anode voltage (volt)
e_f	filament terminal voltage (volt)
e_g	grid voltage (volt)
e_g'	equivalent grid voltage (volt)
E_S	saturating voltage (volt)
E_a	steady anode voltage (volt)
$\mathcal{E}_a$	alternating anode voltage, max. value (volt)
$\mathcal{E}_g$	alternating grid voltage, max. value (volt)
g	mutual conductance (mho)
G	perveance
i	electron current (ampere)
i_a	anode current, instantaneous value (ampere)
i_g	grid current (ampere)
i_f	filament current (ampere)
I_a	anode mean or direct current (ampere)
I_g	grid mean or direct current (ampere)
I_S	saturation current (ampere)
$\mathbf{I}_a$	anode alternating current, fundamental component max. value (ampere)
k	amplification constant
L	life of a cathode (hour)
L_g	ratio of total length of effective grid wires to axial length of the grid.
m	number of points of support of the filament
n	number of grid wires parallel to filament

p	pitch of adjacent grid wires between centers (cm)
P	power output of a triode (watt)
q	portion of the filament power radiated through the anode
r	anode resistance (ohm)
r	mean anode resistance (ohm)
R	a-c impedance (resistive) of the anode circuit (ohm)
r_f	filament radius (cm)
r_g	grid wire radius (cm)
s	number of supports of grid wires which lie in the stream of electrons
t	thickness of grid support (cm)
w	permissible anode loss per unit area (watt/cm^2)
W_a	anode loss (watt)
W_f	filament heating power (watt)
W_g	grid internal loss (watt)
W_e	grid excitation loss (watt)
W_l	grid leak loss (watt)
x_a	distance of anode to cathode axis (cm)
x_g	distance of grid to cathode axis (cm)
z_a	mean shortest distance of anode to filament (cm)
z_g	mean shortest distance of grid to filament (cm)
η	anode power conversion efficiency

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WIRELESS ECHOES OF LONG DELAY*

By

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Summary—*The paper discusses the various possible ways in which long delayed radio echoes may be propagated. It is proved that the echoes cannot be due to waves which are propagated in any of the following manners:*

(1) Along very long paths in the highly ionized part of the earth's atmosphere.

(2) In any part of the earth's atmosphere so highly ionized that the group or signal velocity is very small.

(3) In any part of space outside the earth's atmosphere so highly ionized that the group velocity is very small.

(4) In the non-conductive layer of air between the earth's surface and the lower surface of a highly conductive layer in the higher atmosphere.

In all of these four cases the attenuation will be altogether prohibitive for all echoes with more than 10 seconds retardation.

In case (3) a pure electron atmosphere would not give any attenuation, but it is also proved that such an atmosphere having the necessary density will be so rapidly dispersed by the electrostatic forces that it cannot persist for any appreciable length of time. In the ionization bands which may cause these echoes the total space charge must consequently be very nearly equal to zero. And in this case the attenuation is great.

The long delayed echoes must therefore be due to signal waves having travelled very long distances (about 3.10^5t km, t being the retardation in seconds, either along bounding surfaces between ionized and empty parts of space outside the earth's atmosphere but "within the earth's magnetic field," or in free space, having been subjected to one or more reflections at such boundary surfaces.

The geometry of those ionized layers which causes delays up to 30 (possibly 60) sec. will mainly be controlled by the earth's magnetic field as shown by Störmer.

Beside these echoes originating within the earth's magnetic field the writer has predicted that echoes might occur with delays of several minutes by reflections from ionization-bands outside the influence of the earth's magnetic field. Such long delayed echoes—3 min. 15 sec. and 4 min. 20 sec.—have recently been reported by Jörgen Hals.

INTRODUCTION†

WITH regard to the origin of the long delayed echoes of short radio waves—λ about 31 m—observed by Jörgen Hals, C. Störmer,[1] Balth van der Pol,[2] and others, rather great dif-

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† Footnotes in square brackets are added after the date when this paper was read before the Royal Danish Society of Science.

[1] C. Störmer, *Nature*, **122,** 681, 1928; **123,** 16, 1929. *C. R.* **187,** 811, 1928; *E. N. T.* **5,** 483, 1928. ["*L'Onde Electrique,*" **7,** 531–532, 1928.]

[2] Balth. van der Pol, *Nature* **122,** 878, 1928. ["*L'Onde Electrique,*" **7,** 534–537, 1928.]

ference of opinion prevails between the various authors who have treated the question. Some, as for instance C. Störmer,[1] P. O. Pedersen,[3] and K. W. Wagner,[4,7] assume the echoes to be caused by the radio waves being reflected from or propagated along swarms or bands of electrons out in space while others, as for instance Balth. van der Pol,[2] E. V. Appleton[5] and M. von Ardenne,[6,8] assume the long delay of the echoes to be due to particular conditions existing along the path of the waves in—or bounded by—the ionized part of the earth's atmosphere. As to the manner in which the propagation of the waves should then occur in order to show such great delays—up to about 30 seconds as were at the time observed—these authors have, however, very different opinions. M. von Ardenne[6] thus assumes that the waves simply travel round the earth a sufficient number of times—some hundreds—in the Kennelly layer the attenuation of which he assumes to be so small for short waves that in spite of the great distance travelled they will arrive at the receiver with sufficient strength. Appleton[5] points out that the length of the time of delay depends upon the group velocity of the waves in the medium in question and if this velocity is small then the length of path will also be correspondingly smaller. Appleton[9] further points out, however, that the waves will be too strongly attenuated in the ionized part of the atmosphere—at least at heights less than about 600 km. But he suggests another possibility, namely that the lower boundary of the ionized layer of the upper atmosphere acts as a sharply defined reflecting "shell."[10] Balth. van der Pol[11] also suggests the long echo times to be due to very small group velocities at places where the electrons are so densely crowded that the refractive index for the waves in question approaches zero. But he hardly pays the necessary attention to the attenuation of the waves under such circumstances.

We shall later discuss these various possibilities, but since a detailed treatment of the very complicated conditions of the path is extremely difficult, and since it is hardly possible to give an account of all the geometrically possible tracks of the waves, and, finally, since we know so very little about the atmosphere above 150 km, we shall try to find some criteria of general validity to throw light upon the matter.

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[3] P. O. Pedersen, "Radiofoniens Aarbog 1929," 9–25 (Copenhagen, October, 1928).

[4] K. W. Wagner, *E. N. T.* **5**, 483, 1928.

[5] E. V. Appleton, *Nature* **122**, 879, 1928.

[6] M. v. Ardenne, *Populär Radio*, 345, Copenhagen, December, 1928.

[7] [Deslandres, *L'Onde Electrique*, **7**, 532–3, 1928.]

[8] [H. S. Jelstrup, *L'Onde Electrique*, **7**, 538–540, 1928.]

[9] loc. cit.

[10] [Jelstrup, loc. cit., holds a similar view of the phenomenon.]

[11] loc. cit.

1. The Attenuation of Electromagnetic Waves in a Homogeneous Ionized Atmosphere.

Plane electromagnetic waves propagated in a homogeneous medium are attenuated at the rate

$$e^{-\gamma_0 x + j\omega(t - x/v)} = e^{-\gamma_0 x + j\omega(t - n(x/c))} \tag{1}$$

where γ_0 is the attenuation constant and

$$v = \frac{c}{n} \tag{2}$$

is the phase velocity of the waves; $c = 3.10^{10}$ cm sec^{-1} being the velocity of the waves in empty space and n the refractive index of the medium for waves of the frequency $f = \omega/2\pi$.

The parameters γ_0 and n are determined by [12]

$$\gamma_0 = \frac{\omega}{c}\sqrt{\sqrt{\frac{\epsilon^2}{4} + \left(2\pi c^2 \frac{\sigma}{\omega}\right)^2} - \frac{\epsilon}{2}} \tag{3}$$

and

$$n = \sqrt{\sqrt{\frac{\epsilon^2}{4} + \left(2\pi c^2 \frac{\sigma}{\omega}\right)^2} + \frac{\epsilon}{2}} \tag{4}$$

where ϵ is the dielectric constant in *esu* of the medium and σ its conductibility in *emu*

From (3) and (4) it follows that [13]

$$n\gamma_0 = 2\pi c\sigma. \tag{5}$$

A signal "carried" by a wave of frequency $\omega = 2\pi f$ travels with the velocity u determined by [14]

$$u = \frac{c}{n + \omega \dfrac{dn}{d\omega}} = \frac{nc}{n^2 + \dfrac{1}{2}\omega \cdot \dfrac{d(n^2)}{d\omega}} = \frac{nc}{g};\; g = n^2 + \frac{1}{2}\omega \cdot \frac{d(n^2)}{d\omega}. \tag{6}$$

The attenuation of a wave which has travelled the distance x is according to (1) determined by

$$e^{-\gamma_0 x} \tag{1'}$$

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[12] P. O. Pedersen, "Propagation of Radio Waves," p. 117. (G. E. C. Gad. Copenhagen, 1927.) Cited in the following as *P. R. W.*

[13] *P. R. W.*, p. 118 (6a).

[14] *P. R. W.*, p. 174 (68).

and the attenuation of a signal having travelled t seconds is consequently

$$e^{-\gamma_0 \mu t} \tag{7}$$

since $x = u \cdot t$.

According to (5) and (6) we have

$$\gamma_0 x = \gamma_0 u t = \frac{\gamma_0 n c}{n^2 + \frac{1}{2}\omega \cdot \frac{d(n^2)}{d\omega}} \cdot t = \frac{2\pi c^2 \sigma}{g} \cdot t = \frac{t}{T_0} \tag{8}$$

where

$$\tau_0 = \frac{g}{2\pi c^2 \sigma} \tag{9}$$

is the time the signal wave must travel to have its amplitude decreased in the ratio e^{-1}.

The above formulas are valid generally for homogeneous media in which the attenuation of the waves is not excessively high. If the medium is an ionized atmosphere with n ions (electrons) per cm^3 and these are on the average exposed to ν collisions per second, then the dielectric constant of the medium and its conductibility are determined by [15]

$$\epsilon = 1 - N \cdot 4\pi \frac{e^2}{m} \cdot \frac{1}{\omega^2 + \nu^2} \text{ [esu] and } \sigma = N \cdot \frac{e^2}{mc^2} \cdot \frac{\nu}{\omega^2 + \nu^2} \text{ [emu]} \tag{10}$$

and these formulas are valid in the cases in which we may neglect the influence of the magnetic field of the earth.

By inserting

$$k = N \cdot 2\pi \frac{e^2}{m} = \frac{1}{2}\kappa\omega^2 \quad \text{or} \quad \kappa = N \cdot 4\pi \cdot \frac{e^2}{m\omega^2} \tag{11}$$

the equations (10) will be

$$\epsilon = 1 - \frac{2k}{\omega^2 + \nu^2} = 1 - \kappa \frac{\omega^2}{\omega^2 + \nu^2} \text{ and } 2\pi c^2 \frac{\sigma}{\omega} = \frac{\nu}{\omega} \cdot \frac{k}{\omega^2 + \nu^2} = \frac{\frac{1}{2}\kappa\omega\nu}{\omega^2 + \nu^2}. \tag{10'}$$

From (4) and 10) we then find[16]

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[15] *P. R. W.*, p. 121, (11).
[16] *P. R. W.*, p. 172, (55).

$$g=n^2+\frac{\omega}{2}\cdot\frac{d(n^2)}{d\omega}=\frac{1}{2}\left\{1-\frac{\kappa\omega^2\nu^2}{(\omega^2+\nu^2)^2}+\frac{[(1-\kappa)\omega^2+\nu^2]^2-\left[1-\frac{1}{2}\kappa\right]\frac{\kappa\omega^2\nu^2}{(\omega^2+\nu^2)^2}}{\sqrt{[(1-\kappa)\omega^2+\nu^2]^2+\kappa^2\omega^2\nu^2}}\right\} \tag{12}$$

For $(1-\kappa)\omega^2+\nu^2 \gg \kappa\omega\nu$ or $\kappa \ll \omega^2+\nu^2/\omega^2+\omega\nu$ we get

$$g \cong 1-\frac{1}{2}\cdot\frac{\kappa\omega^2\nu^2}{(\omega^2+\nu^2)^2}\cdot\frac{(2-\frac{3}{2}\kappa)\omega^2+(2-\frac{1}{2}\kappa)\nu^2}{(1-\kappa)\omega^2+\nu^2}\,. \tag{13}$$

If further $\nu \ll \omega$, then we have

$$g \cong 1-\kappa\frac{\nu^2}{\omega^2}\cdot\left(1+\frac{\kappa}{4(1-\kappa)}\right) \tag{13'}$$

which for $\kappa \ll 1$ reduces to

$$g \cong 1-\kappa\frac{\nu^2}{\omega^2}\,. \tag{13''}$$

In all cases where $\kappa \ll \omega^2+\nu^2/\omega^2+\omega\nu$, $g=1$ will consequently give a good approximation.

For $\kappa=2$ we have

$$g=\frac{\nu^4}{(\omega^2+\nu^2)^2} \cong \nu^4 .\ \text{—}^4\omega \tag{13'''}$$

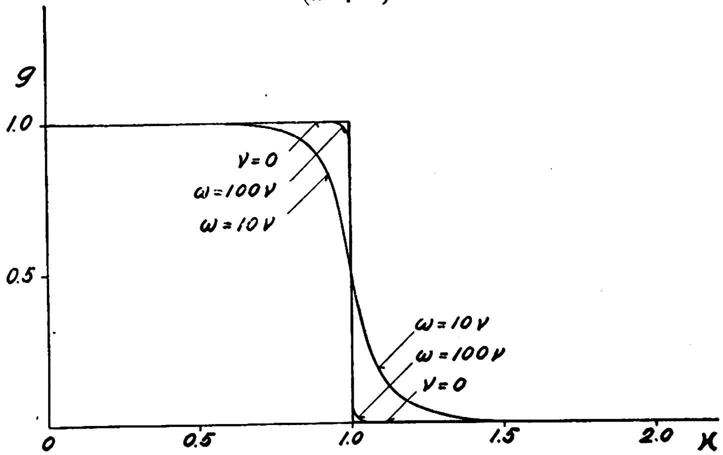

Fig. 1—The factor g as a function of κ.

Fig. 1 shows g as a function of κ and we see that for $\nu \leqq 10^{-2}\cdot\omega$ we get with good approximation $g=1$ for $0<\kappa<1$ while g is very nearly zero for $\kappa>1$. For $\kappa=1$ we get $g \cong 1/2$.

From (6) it appears that for $g=1$ the group velocity is proportional to n. Therefore it will be of some interest to determine the smallest value which n can assume. By means of the equations (4), (10), and (11) we easily deduce that n will have minimum value for $\kappa=2$ and that

$$n_{\min}=\frac{\nu}{\sqrt{\omega^2+\nu^2}}\cong\frac{\nu}{\omega} \tag{14}$$

while the corresponding value of the dielectric constant will be

$$\epsilon=\frac{\nu^2-\omega^2}{\nu^2+\omega^2}\cong-1. \qquad (15)^{17}$$

Another point of particular interest is the one corresponding to $\kappa=(\omega^2+\nu^2)/\omega^2\cong 1$, here we get $n\cong\sqrt{\nu/2\omega}$ and $\epsilon=0$.

The density of electrons [18] corresponding to the various values of κ is according to (11) determined by

$$N=\frac{\kappa\omega^2}{4\pi\frac{e^2}{m}}=3.14\cdot10^{-10}\cdot\kappa\cdot\omega^2 \text{ electrons per cm}^3. \tag{16}$$

The refractive index is thus independently of the air pressure minimum for

$$N=6.29\cdot10^{-10}\cdot\omega^2=2N_0. \tag{17}$$

Assuming that $\nu\ll\omega$ we further get $\epsilon\cong0$ for $N=N_0=3.14\cdot10^{-10}\cdot\omega^2$.

In Table I is given a view of the constants which are of interest for the propagation problem here considered. Besides ϵ, n, and g the table contains the formulas for the group velocity u, for the constant T_0; for the attenuation constant γ_0 and for the attenuation-factor $e^{-\gamma_0\lambda}$ which latter indicates the attenuation of the amplitude by the

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[17] In the deduction of (14) and (15) no account is taken of the influence of the earth's magnetic field on the conductibility and on the dielectric constant. If we do so then we get (see *P. R. W.*, p. 122) that for

$$\kappa=2\frac{\omega-h}{\omega} \text{ we get } n_{\mathrm{I}\ \min}\ \frac{\nu}{\omega-h}$$

and for

$$\kappa=2\frac{\omega+h}{\omega} \text{ we get } n_{\mathrm{II}\ \min}\cong\frac{\nu}{\omega+h}$$

[18] Generally we speak only of electrons because one electron will influence the propagation of the waves about as much as 50,000 ions.

wave travelling a distance equal to one wavelength in the medium in question.

TABLE I

The values given are valid for $\nu < < \omega$; $N_0 = 3.14 \cdot 10^{-10} \cdot \omega^2$.

$\kappa = \frac{N}{N_0}$	ϵ	n	g	u	T_0	γ_0	$e^{-\gamma_0 \lambda}$
$0 \leqq \kappa < 1$	$1-\kappa$	$\sqrt{1-\kappa}$	1	$c\sqrt{1-\kappa}$	$\frac{2}{\kappa\nu}$	$\frac{\nu}{2c} \cdot \sqrt{\frac{\kappa^2}{1-\kappa}}$	$e^{-\pi \kappa\nu/(1-\kappa)\omega}$
$\kappa = 1$	0	$\sqrt{\frac{\nu}{2\omega}}$	$\frac{1}{2}$	$c\sqrt{2\frac{\nu}{\omega}}$	$\frac{1}{\nu}$	$\frac{1}{c} \cdot \sqrt{\frac{\omega\nu}{2}}$	$e^{-2\pi}$
$\kappa = 2$	-1	$n_{min} = \frac{\nu}{\omega}$	$\frac{\nu^4}{\omega^4}$	$\left(c \cdot \frac{\omega^3}{\nu^3}\right)^*$	$\left(\frac{\nu^3}{\omega^4}\right)^*$	$\frac{\omega}{c}$	$e^{-2\pi(\omega/\nu)}$

* These expressions are formally in agreement with formulas (6) and (9), but have no physical significance. See the text.

From the table it appears that for $\kappa = 2$ we find $u > c$. Since, however, we must always necessarily have $u \leqq c$ the method of determining the group velocity applied has led to wrong results, and the corresponding value of T_0 is consequently also of no significance. To emphasize these circumstances both of these expressions are placed in brackets.

The reason we come to such unreasonable results at great densities of electrons—or more correctly, at great values of κ—is quite evident. In the deduction of the formula (6) for the group velocity we assumed the "signal" to be produced by the superposition of a number of continuous waves having a frequency only very slightly different from the carrier frequency $f = \omega/2\pi$. From the last column in the table it appears, however, that for $\kappa = 2$ the attenuation will be so great that the waves will lose entirely their periodical character. If for example we put $\omega = 6.10^7$ and $\nu = 6.10^3$, then by travelling one wavelength the amplitude will have decreased at the rate $e^{-2\pi.10^4}$, and in this case the formula (6) must consequently lead to unreasonable results. There is, however, no reason to enter further into this question since it is quite evident that the attenuation here is so great that waves entering a medium of the character considered within a period will be reduced practically to zero.

Even for $\kappa = 1$ where we get $u < c$, it is doubtful whether the values found for u and T_0 will be correct, since in this case the attenuation over a wavelength will be $e^{-2\pi}$. The long delayed echoes observed would consequently be unable to pass through a medium corresponding to $\kappa = 1$, and in the following we shall therefore confine ourselves to treat the transmission of signals through such media for which $0 \leqq \kappa < 1$. For these we get

$$T_0 = \frac{2}{\kappa\nu} \tag{18}$$

and after t seconds the attenuation of plane waves in such a medium will be

$$e^{-t/T_0} = e^{-\kappa\nu/2 \cdot t}. \qquad (19)^{19}$$

In order to get in t sec. a track length travelled by the wave of ξct instead of ct as would be the case in empty space we must have

$$\xi = \sqrt{1-\kappa} \quad \text{or} \quad \kappa = 1-\xi^2. \tag{20}$$

During t sec. the amplitude will decrease at the rate

$$e^{-t/T_0} = e^{-\kappa\nu t/2} = e^{-\frac{1}{2}(1-\xi^2)\nu t}. \tag{21}$$

If the track traversed by the wave is to be essentially shorter than what it would be during an equal time in empty space then must $\xi^2 \ll 1$. In that case the attenuation S during t sec. will be approximately

$$S = e^{-1/2\nu t} \tag{22}$$

which is exactly the expression found by Appleton.

To $\xi \rightarrow 0$ corresponds $\kappa \rightarrow 1$ and the density of electrons $N \rightarrow N_0 = 3.14 \cdot 10^{-10} \cdot \omega^2$.

Appleton assumes that for a height of 250 km above the surface of the earth we may put $\nu = 1000$. I should be inclined for this height to apply $\nu = 360$.* For $t = 10$ sec. the attenuation factors would then be respectively

$$S = e^{-5000} \quad \text{and} \quad S = e^{-1800}$$

and accordingly we would get an attenuation much too great even for delay of 10 sec. by applying even the value for ν assumed by me. For a height of 400 km Appleton assumes $S = e^{-50}$ for $t = 10$ sec. which he, however, considers as too great an attenuation; nevertheless he adds: "But if there were sufficient ionization at heights of 600 km or more, it is certain that retardation without much absorption could take place, although our inadequate knowledge of the values of ν for such regions precludes a more quantitative statement."

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[19] For $\kappa \rightarrow 1$ the attenuation deduced here is in agreement with the one given by E. V. Appleton (loc. cit.).

* *P. R. W.* Appendix, p. 6, Fig. IX, 1; compare also Table III below.

In estimating the value of ν, only the collisions of the electrons with neutral molecules have, however, so far been considered, and not their collisions with other electrons or ions. The error thus introduced is small as long as the number of electrons and ions are exceedingly small in comparison to the number of neutral molecules, which will always be the case— at least for the circumstances here considered —as long as the air pressure is not extremely small. The error will, however, be very great if the air pressure approaches zero, or if the number of electrons plus ions approaches—or even exceeds—the number of neutral molecules.

We shall, therefore, now consider the free path l and the number of collisions ν of an atmosphere having solely N electrons per cm³. According to W. Schottky and H. Rothe[20] the mean free path is determined by

$$l = \frac{1}{\pi\sqrt{2} \cdot Nd^2} \tag{23}$$

where d is the "diameter" of the electron. The latter is determined by the potential electric energy of the two colliding electrons—or of an electron and a positive (or negative) ion—being equal to $1/\eta$ of the kinetic energy of a single electron, when they are at a distance d from each other.

The mean value of the kinetic energy of a single electron will, at a temperature $T^{0\text{abs}}$, be determined by

$$\frac{1}{2}mU^2 = \frac{3}{2}kT = 2.058 \cdot 10^{-16}T, \tag{24}$$

where $k = 1.372 \cdot 10^{-16}$ is the Boltzmann constant.

The number of collisions ν is determined by

$$\nu = \frac{U}{l} = \pi\sqrt{2}Nd^2\sqrt{\frac{3kT}{m}} \tag{25}$$

the mean velocity of the electron—according to (24)— being

$$U = \sqrt{\frac{3kT}{m}} = 6.76 \cdot 10^5\sqrt{T}. \tag{25'}$$

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[20] *Handb. d. Experimentalphysik,* **XIII**, 2, 41, 1928.

The diameter d will then be determined by

$$\frac{e^2}{d}=\frac{1}{\eta}\cdot\frac{1}{2}mU^2=\frac{1}{\eta}\cdot\frac{3}{2}kT=\frac{1}{\eta}\cdot 2.058\cdot 10^{-16}T \tag{26}$$

or

$$d=\frac{\eta e^2}{\frac{3}{2}kT}=1.107\cdot 10^{-3}\frac{\eta}{T}. \tag{26$'$}$$

Schottky and Rothe assume as an upper limit $\eta=100$ to which corresponds

$$d=0.1107\cdot\frac{1}{T}\cdot[\text{cm}]. \tag{27}$$

By (23) and (26′) the free path will generally be

$$l=1.84\cdot 10^5\frac{T^2}{\eta^2 N} \tag{28}$$

and the number of collisions

$$\nu=3.68\cdot\frac{\eta^2}{T^{3/2}}\cdot N. \tag{29}$$

At for instance $T=300$ deg. abs. we then get $\nu=7.08\cdot 10^{-4}\eta^2 N$.

The value $\eta=100$ assumed by Schottky and Rothe[21] for other purposes is no doubt too great for the object here considered. $\eta=1$ is, on the other hand, no doubt too small. The formula (29) shows, however, the necessity of fixing somewhat narrower limits for η or for the diameter d which depends on η.

For reasons which will be given in the following we shall, however, not consider collisions between two electrons but between an electron and a positive—or negative—ion, and we assume the latter to be at rest as well before as after the collision.

The angle of deflection, 2θ, of the electron (see Fig. 2) is determined by[22]

$$\operatorname{tg}\theta=\frac{e^2}{mU^2p} \tag{30}$$

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[21] loc. cit.

[22] See, for example, *Handb. d. Physik,* **XXIV**, p. 4, 1927.

where U is the velocity of the electron and p the distance of the ion from the straight path of the electron.

By means of (25′) and (26) equation (30) is reduced to

$$\operatorname{tg}\theta = \frac{1}{2\eta}\cdot\frac{d}{p} = \frac{\alpha}{p} \tag{31}$$

where

$$\alpha = \frac{d}{2\eta} = \frac{e^2}{3kT} = \frac{5.5\cdot 10^{-4}}{T}. \tag{31'}$$

The velocity component of the electron after the collision in the original direction will be equal to the original velocity multiplied by

$$\xi' = 1 - 2\sin^2\theta = \frac{p^2-\alpha^2}{p^2+\alpha^2}. \tag{32}$$

That fraction of the original velocity which is deflected from the original direction is consequently

$$\xi = 1 - \xi' = \frac{2\alpha^2}{p^2+\alpha^2}. \tag{32'}$$

By integrating this quantity over a circular area at right angles to the original direction of the electron, and having the radius p_0, we get

$$A = \int_0^{p_0} \xi\pi\cdot 2\pi p\,dp = 2\pi\alpha^2\cdot \lg\frac{p_0{}^2+\alpha^2}{\alpha^2}. \tag{33}$$

If we put

$$A = \pi d_0{}^2 \tag{34}$$

then d_0 may be taken as a kind of equivalent diameter and we get

$$d_0 = \alpha\sqrt{2\lg\frac{p_0{}^2+\alpha^2}{\alpha^2}}. \tag{35}$$

Table II contains some values of d_0/α for various values of p_0/α. d_0 increases steadily but slowly with increasing values of p_0.

TABLE II

Values of $\sqrt{2lg\frac{p_0{}^2+\alpha^2}{\alpha^2}}$ for various values of $\frac{p_0}{\alpha}$.

$\frac{p_0}{\alpha}$	1	2	3	4	5	10	20	100	1000
$\sqrt{2lg\frac{p_0{}^2+\alpha^2}{\alpha^2}}=\frac{d_0}{\alpha}$	1.18	1.79	2.15	2.38	2.55	3.04	3.46	4.29	5.26

These calculations assume, however, that along the entire curved part of the path the colliding particles are so far away from other charged particles that their movements are practically unaffected by these. Since the density of electrons, according to Table I shall be $N=N_0=3.14\cdot 10^{-10}\ (2\pi)^2\cdot 10^{14}=1.24\cdot 10^6$ electrons per cm³ in order to reduce to a considerable extent the group velocity of the 30-m wave, and since according to (31′) $\alpha=10^{-6}$ cm for $T=550°$ abs. it will hardly be justifiable to assign to p_0/α a value much higher than 100 to which corresponds $d_0=4.29\alpha$. On the other hand, the value p_0/α should no doubt be higher than 3 to which corresponds $d=2.15\alpha$. Since for the following it is not essential to know the exact value of d_0 we simply put

$$d_0=3\alpha=\frac{1.65\cdot 10^{-3}}{T}\ [\mathrm{cm}] \tag{35'}$$

which is 67 times smaller than the one assumed by Schottky and Rothe.

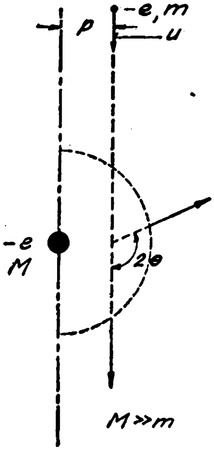

Fig. 2—Collision between an electron and a fixed negative ion.

To (35′) corresponds a length of free path

$$l=8.3\cdot 10^4\frac{T^2}{N} \tag{36}$$

and a number of collisions

$$\nu=\frac{U}{l}=\frac{6.7\cdot 10^5 T^{1/2}}{8.3\cdot 10^4 T^2}\cdot N=8.1\cdot\frac{N}{T^{3/2}}. \tag{37}$$

Table III shows for some values of T the corresponding values of the ratio ν/N.

TABLE III

The ratio ν/N calculated by means of (37). Values of ν for $N=1.2\cdot 10.^6$

$T=$	100°	400°	900°	1600°	2500°	3600°	10000°	250.000° abs.
$\frac{\nu}{N}=$	$8.1\cdot 10^{-3}$	$1.0\cdot 10^{-3}$	$3.0\cdot 10^{-4}$	$1.3\cdot 10^{-4}$	$6.5\cdot 10^{-5}$	$3.8\cdot 10^{-6}$	$8.1\cdot 10^{-5}$	$6.5\cdot 10^{-8}$
$N=1.2\cdot 10^5;\ \nu=$	$1\cdot 10^4$	$1.2\cdot 10^3$	$3.6\cdot 10^2$	$1.6\cdot 10^2$	78	46	9.7	$7.8\cdot 10^{-2}$

We cannot, however, insert the values of ν thus obtained in the formula (22) and by this means calculate the attenuation of the waves in an atmosphere consisting only of electrons. In this case where the colliding particles are perfectly identical a collision will exert no influence at all on the propagation of the waves since we have

$$\begin{gathered} v_{1x}+v_{2x}=v_{1x}'+v_{2x}'\,;\quad v_{1y}+v_{2y}=v_{1y}'+v_{2y}'\,;\quad v_{1z}+v_{2z}=v_{1z}'+v_{2z}'\,; \\ \text{and} \qquad v_1^2+v_2^2=v_1'^2+v_2'^2, \end{gathered} \tag{38}$$

where v_1 and v_2 are the velocities of the electrons before and v_1' and v_2' their velocities after the collision.

The sum of the "current" components, viz. $-e\cdot(v_{1x}+v_{2x})$, $-e\cdot(v_{1y}+v_{2y})$, $-e\cdot(v_{1z}+v_{2z})$, and of the kinetic energy, viz. $\frac{1}{2}m\cdot(v^2_1+v_2^2)$, of two electrons is thus not altered. *Radio waves are consequently propagated in an atmosphere consisting solely of electrons without suffering any loss and in the same manner as if there were no collisions at all between the electrons.*[23]

On account of their unstable character such pure electron "atmospheres" ("layers" or "bands") can not, however, play any important rôle in the cases of propagation of radio waves here considered. To illustrate this we will consider such a large, plane band of electrons or ions. The mutual repulsion will cause the thickness of the band to increase and consequently the density of electrons to decrease.

It is easily proved that the density q of the charge within a plane layer, see Fig. 3, part I, decreases with increasing time according to the following formula

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[23] [In a letter in *Nature* (February 2, 1929, p. 166) L. H. Thomas has also called attention to the influence which the other electrons and ions, which are present in the neighbourhood of an electron or ion, exert on the length of the effective mean free path of this electron or ion. He does not, however, mention the fundamental difference, with regard to the attenuation of radio waves, between collisions taking place between charged particles of equal charges and equal masses and collisions in which the two particles have unequal masses or charges (or both).]

$$q_t = \frac{q_0}{1+2\pi q_0 \cdot \dfrac{e}{m} \cdot t^2}, \quad [\text{esu}] \qquad (39)$$

q_0 being the density at the time $t=0$ and q_t the density t seconds later.

Fig. 3, part II, sections, 1, 2, and 3 show the dispersion of a plane homogeneous layer having at the time $t=0$ a density of $N=1\cdot 10^5$ electrons per cm^3. Section 2 shows the distribution $1\cdot 10^{-7}$ sec. and section 3 shows $2\cdot 10^{-7}$ sec. later.

Fig. 3 *ABCD* represent parts of large, plane bands of electrons or ions. In part I q_0 is the density of the charge at the time $t=0$.

From this it appears that a band of electrons having a density of 10^6 can exist only if inside the band there is a positive space charge of practically equal value, and evidently it must be mainly the less mobile positive particles which determine the geometric relations of the band.

If the positive space charge consists of mono-valent positive ions then the number of "effective" collisions ν is determined simply by means of (37), N being the density of the positive and negative atomic or molecular ions, and the attenuation of the wave is found by inserting in (22) the value of ν thus obtained. If the positive and negative ions are poly-valent, then ν and accordingly also the attenuation will be even greater.

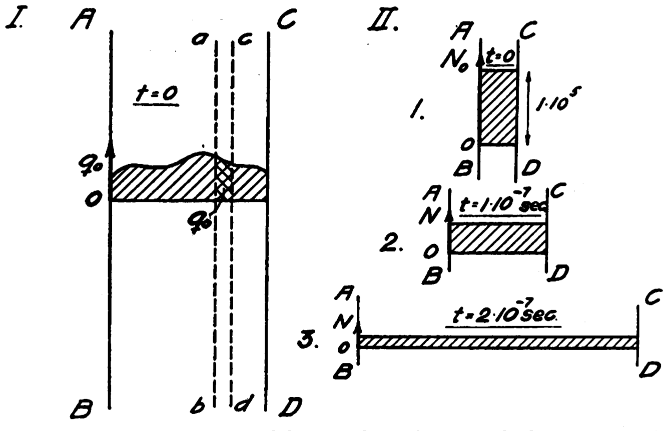

Fig. 3—*ABCD* represent parts of large plane bands of electrons or ions. In part I q_0 is the density of the charge at the time $t=0$.

In order that a 30-m wave shall travel at a slow group velocity the electron density of the medium must, as is shown in the preceding, be about $1.2\cdot 10^6$ electrons per cm^3. For a temperature of 900 deg. abs.

according to Table III we then get $\nu=360$. A wave travelling in such a medium for 10 seconds will, according to (22), suffer an attenuation $S=e^{-1/2\nu t}=e^{-1800}$. For $T=1600$ deg. abs. we get $S=e^{-800}$.

From the preceding it appears that even if we pay no attention to the attenuation suffered by the waves in consequence of collisions between electrons and neutral air molecules, the radio waves will, nevertheless, in consequence of the collisions between the electrons and the necessarily existing positive ions be so greatly attenuated that the long delayed wireless echoes cannot have travelled a considerable part of the time within such an electron-ion-atmosphere.[24]

2. The Propagation of Electromagnetic Waves Along the Boundary Surface between an Ionized and a Non-Ionized Part of the Earth's Atmosphere.

Having now shown that the long delayed echoes cannot be due to the propagation of the waves *in* a space containing so many electrons that the group velocity is many times smaller than the velocity of the waves in free space, because in such a medium the attenuation will be too great, it may be appropriate to investigate whether the long delays may be due to the fact that the waves are propagated along the boundary between such a medium and either empty space or an non-ionized air space. A smaller attenuation would be possible in either case, since the loss of energy in the ionized part could be compensated for by the field energy in the non-ionized part. Something similar occurs in a very pronounced manner in the propagation of the waves along the surface of the earth.[25] We shall therefore consider now a little more closely the propagation of the waves along such a plane boundary surface.

The boundary surfaces will, of course, generally be curved, but the radii of curvature must at all events be very great if propagation along such surfaces shall be of any significance at all for the cases here considered. Further the concavity of the boundary surface must face toward the earth and toward the non-ionized part of space, and will thus serve to collect the radiation-energy. The border between the conductive and the non-conductive part of space will never be perfectly sharp because some electrons and ions will penetrate into the otherwise nonconductive part. Under such conditions the waves will be refracted and show a tendency to follow the curved boundary surface.

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[24] Even if the long delayed echoes cannot be due to propagation in media having a very small group velocity, such propagation may play an important rôle in the various methods of determining the altitude of the highest point of the path of radio rays, often called the height of the ionized layer, and may have caused some of the discrepancies between the values which various experimenters have found for this altitude, insofar as the discrepancies are not due to "reflections" from various "layers." This question is treated of in *P. R. W.*, pp. 171–178, 209–212, and the theory is further developed by E. V. Appleton (*Proc.* Phys. Soc., **41**, 43–56, 1928) and J. C. Schelleng (Proc. I. R. E., **16**, 1471–76; November, 1928), whose papers will be a great help to the investigation of this problem in the future.

[25] *P. R. W.*, chap. III, p. 22 and 29–32.

The conditions existing at a plane boundary surface will, no doubt, give an approximately correct representation of the actual conditions. Considerable mathematical difficulties would also be met with in the treatment of a propagation along a curved surface, while the problem of propagation along a plane boundary surface, as shown by J. Zenneck[26], may be solved in a simple manner.

By considering the boundary surface to be an XY-plane and then assuming the plane waves to travel along the plane in the positive direction of the x-axis (see Fig. 4) attenuation along the boundary surface will be determined by

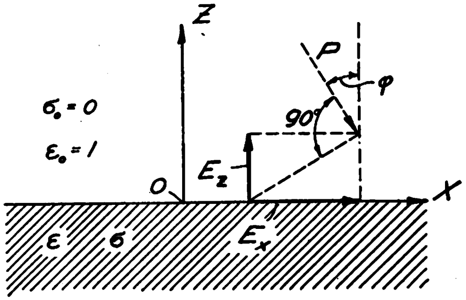

Fig. 4—Propagation of waves along the boundary surface XY. In the upper medium the energy is propagated in the direction indicated by the arrow P.

$$e^{-\gamma_0 x} \tag{1}$$

If we call the phase velocity along the surface v then γ_0 and v can be determined by the following equation[27]

$$\gamma_0 + j\frac{\omega}{v} = j\frac{\omega}{c}\sqrt{\frac{1+j\frac{\omega}{\sigma}\cdot\frac{\epsilon}{4\pi c^2}}{1+j\frac{\omega}{\sigma}\cdot\frac{1+\epsilon}{4\pi c^2}}} \tag{2}$$

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[26] J. Zenneck, *Ann. d. Phys.* (4) **23,** 846, 1907. *P. R. W.*, chap. III.

[27] *P. R. W.*, p. 17 (the symbols used are different).

provided the medium (0) has the dielectric constant $\epsilon_0=1$ and the conductivity $\sigma_0=0$, while the corresponding constants for the ionized medium are ϵ and σ [emu.].

Let the X and Z components of the electric field at the boundary surface be E_x and E_z; then we have

$$\left|\frac{E_x}{E_z}\right|_{z=+0}=\sqrt{\frac{j\frac{\omega}{\sigma}\cdot\frac{1}{4\pi c^2}}{1+j\frac{\omega}{\sigma}\cdot\frac{\epsilon}{4\pi c^2}}} \tag{3}$$

where the index $z=+0$ indicates that the field components shall be taken in the insulated medium but immediately at boundary surface $z=0$.

If $0<\epsilon<1$ and $\omega/\sigma\cdot\epsilon/4\pi c^2\gg1$ then (2) and (3) will give

$$\gamma_0=\frac{2\pi c\sigma}{(1+\epsilon)\sqrt{\epsilon(1+\epsilon)}},\quad v=c\sqrt{\frac{1+\epsilon}{\epsilon}}=u \text{ and } \left|\frac{E_x}{E_z}\right|_{z=+0}=\sqrt{\frac{1}{\epsilon}} \tag{4}$$

Since here there is no dispersion, the group velocity is equal to the phase velocity, and both are greater than the velocity c in empty space. This is quite natural, since the energy propagation does not occur along the boundary surface but in a direction P (see Fig. 4) which is perpendicular to the resultant of E_x and E_z. If ϵ is extremely small, then the energy propagation will take place practically perpendicular to the boundary surface.

For the attenuation exponent we get (compare equation (10′) in sect. 1)

$$\gamma_0x=\gamma_0ut=\frac{2\pi c^2\sigma}{\epsilon(1+\epsilon)}\cdot t=\frac{\frac{1}{2}\kappa\frac{\omega^2\nu}{\omega^2+\nu^2}}{\left(1-\kappa\frac{\omega^2}{\omega^2+\nu^2}\right)\left(\cdot2-\kappa\frac{\omega^2}{\omega^2+\nu^2}\right)}\cdot t\cong\frac{\frac{1}{2}\kappa\nu t}{(1-\kappa)(2-\kappa)} \tag{5}$$

where the last term is correct only for $\nu\ll\omega$.

In order to obtain a small attenuation exponent the smallest possible value of κ should be chosen. If $\kappa\ll1$ then for (5) we get

$$\gamma_0x=\gamma_0ut=\tfrac{1}{4}\kappa\nu t. \tag{5′}$$

We will later come back to the application of this result.

Next we will consider the case where the density of electrons is so great that $\kappa > 2$, and we therefore have

$$\epsilon = 1 - \kappa \qquad \frac{\omega^2}{\omega^2+\nu^2} < 0.$$

We then get

$$\frac{\omega}{\sigma} \cdot \frac{\epsilon}{4\pi c^2} = \frac{\omega^2+\nu^2}{\kappa\omega\nu} - \frac{\omega}{\nu} \text{ and } \frac{4\pi c^2 \sigma}{\omega} = \kappa \frac{\omega\nu}{\omega^2+\nu^2}.$$

For $\nu \ll \omega$ by means of (2) we get

$$\gamma_0 = \frac{\nu}{2c} \cdot \frac{\kappa}{(\kappa-2)^{3/2} \cdot (\kappa-1)^{1/2}} \text{ and } v = c\sqrt{\frac{\kappa-2}{\kappa-1}} = u. \tag{6}$$

The attenuation exponent will then be

$$\gamma_0 x = \gamma_0 u t = \frac{\kappa\nu}{2(\kappa-1)(\kappa-2)} \cdot t \cong \frac{\nu}{2\kappa} \cdot t \tag{6'}$$

where the last term holds good for $\kappa \gg 1$.

Finally, if $\nu \gg \omega$ then

$$\gamma_0 = \frac{\kappa\omega^2}{4\sqrt{2} \cdot c\nu}, \quad v = c\sqrt{2} = u \tag{7}$$

to which corresponds

$$\gamma_0 x = \gamma_0 u t = \frac{\kappa}{4} \cdot \frac{\omega^2}{\nu} \cdot t. \tag{7'}$$

For a judgment of the conditions we must know how rapidly the amplitude of the waves decrease in the non-conductive medium and in a direction perpendicular to the boundary surface.

Using the symbols from Fig 4 this decrease is determined by

$$e^{-r_0 z}, \tag{8}$$

where r_0 and the phase velocity of the waves w in the direction of the Z-axis is determined by [28]

$$r_0 + j\frac{\omega}{w} = j\frac{\omega}{c} \sqrt{\frac{j\dfrac{\omega}{\sigma} \cdot \dfrac{1}{4\pi c^2}}{1 + j\dfrac{\omega}{\sigma} \cdot \dfrac{1+\epsilon}{4\pi c^2}}}. \tag{9}$$

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[28] *P. R. W.*, p. 17.

TABLE IV.

The attenuation-exponents for the Propagation of Plane Waves along a Boundary Surface between an Ionized and a Non-ionized Part of Space.

Assumptions			Attenuation exponent $\gamma_0 x = \gamma_0 u t$		Group-velocity	Attenuation-constant	$\left\lvert \frac{E_x}{E_z} \right\rvert_{z \to +0}$	$\phi = \text{arctg} \left\lvert \frac{E_z}{E_x} \right\rvert$ $= \tan^{-1} \left\lvert \frac{E_z}{E_x} \right\rvert$
$\epsilon_0 = 1;\ \sigma_0 = 0$			ν	$\nu = \kappa \nu_1$	u	r_0		
$0 < \epsilon < 1$	$\nu << \omega$	$0 < \kappa < 1$	$\frac{\kappa \nu}{2(1-\kappa)(2-\kappa)} \cdot t$	$\frac{\kappa^2 \nu_1}{2(1-\kappa)(2-\kappa)} \cdot t$	$c\sqrt{\frac{1+\epsilon}{\epsilon}}$	$\frac{\kappa \nu}{2c(2-\kappa)^{3/2}}$	$\sqrt{\frac{1}{1-\kappa}}$	$\text{arctg}\sqrt{1-\kappa}$ $= \tan^{-1}\sqrt{1-\kappa} < 45°$
	$(\epsilon \cong 1 - \kappa)$	$0 < \kappa << 1$	$\frac{1}{4} \kappa \nu t$	$\frac{1}{4} \kappa^2 \nu_1 t$	$c\sqrt{\frac{1+\epsilon}{\epsilon}} \cong c\sqrt{2}$	$\frac{\kappa \nu}{4\sqrt{2} \cdot c}$	1	45°
$\epsilon < -1$	$\nu << \omega$	$\kappa >> 2$	$\frac{\nu}{2\kappa} \cdot t$	$\frac{\nu_1}{2} \cdot t$	$\sqrt{\frac{\kappa - 2}{\kappa - 1}}$			
	$\nu >> \omega$	$\kappa > 2$	$\frac{\kappa \omega^2}{4\nu} \cdot t$	$\frac{\omega^2}{4\nu_1} \cdot t$	$c\sqrt{2}$			

For $\kappa < 1$ and $\nu \ll \omega$ we consequently get

$$r_0 = \frac{\kappa\nu}{2c(2-\kappa)^{3/2}}. \tag{10}$$

The values of r_0 and of the ratio $\left|\frac{E_x}{E_z}\right|_{(z=+0)}$ and further of $\phi = \text{arctg} \left|\frac{E_x}{E_z}\right|$ as well as the above obtained results are shown in Table IV.

From this table it appears that the attenuation becomes very great for $\epsilon < -1$ $(\kappa > 2)$ both for $\nu \ll \omega$ and still more for $\nu \gg \omega$. In both cases it is so great that it is out of question that the corresponding wave propagation can form the basis of the long delayed echoes. A closer consideration further shows that the attenuation exponent becomes very great within the interval $0 > \epsilon > -1$ so that this assumption also is out of question.

We then only have the interval $0 < \epsilon < 1 (1 > \kappa > 0)$. For $\kappa \rightarrow 1$ and consequently $\epsilon \rightarrow 0$ we have $\gamma_0 ut \rightarrow \infty$ while for $\kappa = 1/2$ and $\epsilon = 1/2$ the attenuation-exponent $\gamma_0 ut = \nu_1/6 \cdot t$. Taking $\nu_1 = 360$ and $t = 10$ sec. we get an attenuation-factor $S = e^{-600}$ while $\kappa = 1/4$ and $\epsilon = 3/4$ gives $S = e^{-85.7}$ for $t = 10$ sec. In order to obtain by the waves propagated in this manner the long delayed echoes we must consequently have $0 < \kappa < 1/4$ and accordingly $3/4 < \epsilon < 1$.

From Table IV we learn that the attenuation-exponent $\gamma_0 ut$ along the boundary surface will be the smaller, the smaller we have the value of κ. We cannot, on the other hand, have $\kappa = 0$, as in that case we should have no boundary surface at all and therefore the entire energy of the waves would disappear out into space. We are unable to fix any definite limit for the value of κ. However, if the waves must be able to travel several hundred times round the earth, which is necessary in order to obtain echoes up to 10 sec. or even more, then the refractive index $n = 1 - \Delta n$ of the ionized layer must assume such small values that waves leaving the transmitter in a horizontal direction are totally reflected at the boundary surface. At a smaller value of the reduction Δn of the refractive index practically all of the energy radiated would proceed out into space and get lost. We therefore must have

$$\Delta n \geqq \frac{h}{R} \tag{11}$$

where h is the height of the boundary surface above the earth and R the radius of the earth.

Since

$$n = 1 - \Delta n \cong \sqrt{\epsilon} \cong \sqrt{1-\kappa} \cong 1 - \tfrac{1}{2}\kappa$$

we get

$$\kappa = 2\Delta n \geqq 2\frac{h}{R}\,. \tag{12}$$

Taking $h = 160$ km and $R = 6400$ km we find by means of (11) and (12) that $\kappa \geqq 1/20$ and the corresponding value of the attenuation exponent for $t = 10$ sec. is, according to Table IV

$$\gamma_0 ut = \tfrac{1}{4}\kappa^2 \nu_1 t = 2.25 \tag{12'}$$

for $\nu_1 = 360$. The attenuation factor will then be $S = e^{-2 \cdot 25}$.

This attenuation is so small that it will be of no importance in the case here considered; and therefore such a manner of propagation would appear to be a solution of the question and may be so for ionization bands in free space, see sect. 4 below. We shall, however, immediately show that this is not the case for a wave travelling some hundreds of times round the earth. To realize this we must, however, consider the field energy relations in the non-conductive medium during the wave propagation in question. According to Table IV the attenuation constant r_0 (compare (8) and (10)) for $\kappa = 1/20$ and for $\nu_1 = 360$ is

$$r_0 = \frac{\kappa^2 \nu_1^2}{4\sqrt{2} \cdot c} = 5.3 \cdot 10^{-12} [\text{cm}^{-1}] = 5.3 \cdot 10^{-7} [\text{km}^{-1}]$$

and for $\kappa = 1/4$

$$r_0 = 1.3 \cdot 10^{-5} [\text{km}^{-1}].$$

In the derivation of Zenneck's formulas for wave propagation along a plane surface, the non-conductive space is assumed to be of infinite height perpendicular to the boundary surface. The total electromagnetic field energy in this space is equal to the field energy of a layer of the height H_0 with a uniformly distributed field having all over the same intensity as that found at the boundary-surface, where H_0 is determined by

$$H_0 = \int_0^\infty e^{-2r_0 z} \cdot dz = \frac{1}{2r_0}\,. \tag{13}$$

For $\kappa = 1/20$ we have $H_0 = 9.4 \cdot 10^5$ km, and for $\kappa = 1/4$ we have $H_0 = 3.8 \cdot 10^4$ km. If we assume the boundary surface to be at a height of 160 km above the earth the total energy will be respectively $940 \cdot 10^3/160 = 5900$ and $38 \cdot 10^3/160 = 240$ times smaller than assumed. The effective attenuation will, therefore, be very much greater than the value 2.25 according to (12′).

The conditions are rather complicated, partly owing to the fact that the boundary surface is not plane but spherical, and partly owing to the presence of the earth's surface. We cannot therefore directly infer that the above found values of the attenuation exponents, namely $\gamma_0 u t_0 = 85.7$ and 2.25 shall be multiplied respectively by $1/2 \cdot 240 = 120$ and $1/2 \cdot 5900 = 2950$ which would raise their values to respectively 10280 and 6700. The correct figures are, however, not essential in this connection but only the fact, which we can easily derive, that the attenuation with this manner of propagation would be so great that a wave propagation subject to such conditions cannot form the basis of the long delayed echoes.

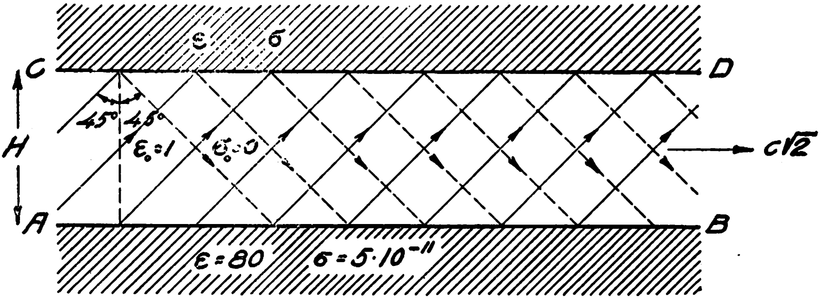

Fig. 5—Schematical representation of the propagation of radio waves between the surface *A-B* of the earth and an ionized layer *C-D*. Assuming the height *H* to be infinitely great, the waves would travel in the direction shown in full drawn lines forming an angle of incidence $\kappa \cong 45$ deg. at a very small value of κ.

If the waves were propagated in that manner then the longer waves as for instance 60 and 90 m would at all events be propagated with a smaller attenuation and consequently give longer delays. Further, echo should generally be obtained after each passage round the earth. Both of these consequences are contradictory to experience and we consequently consider it as impossible that the long delayed echoes are produced in that manner.

This result is also verified by a consideration of Fig. 5 where the full drawn lines show the direction of the energy propagation under the assumption that within the limited layer between the earth's surface $A-B$ and the boundary surface $C-D$ the propagation occurs in the same manner as in unlimited space. The actual propagation will evidently suffer very great losses partly by reflection (at an angle of incidence of about 45 deg.[29] from the ionized layer, partly by reflection of the wave from, and its propagation along, the surface of the earth.

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[29] Compare *P. R. W.*, p. 140.

The most favorable case would be if the propagation occurred between a perfectly even ocean-surface and a perfectly conductive outer shell at a height $h(\cong 160$ km). If we further assume the wave energy to be uniformly distributed between the earth's surface and the outer shell then at $\omega = 6 \cdot 10^7 (\lambda \cong 30m)$ we should get an attenuation constant γ_0 with a value of [30]

$$\gamma_0 \cong \frac{5 \cdot 10^{-3}}{h} \cong 3.1 \cdot 10^{-5} [\mathrm{km}^{-1}].$$

For $x = 3 \cdot 10^6$ km corresponding to an echo period of 10 sec. we get $\gamma_0 x = 93$. The actual attenuation exponent will, no doubt, be considerably greater.

A propagation such as this would only attain a sufficiently low attenuation if the height h were very great—at all events so great that the ionized layer should be located entirely outside the earth's atmosphere.[31]

On the other hand, a propagation in the manner here considered, and repeated reflection from the surface of the ionized layer contribute in a great measure to the remarkably efficient transmission of short waves in the earth's atmosphere which we know from experience. We have in fact that while the direction of the wave-ray in the non-conductive medium forms the angle $\psi = 90° - \phi$ with the boundary-surface, where

$$\mathrm{tg}\,\psi_0 = \left|\frac{E_x}{E_z}\right|_{z=+0} \cong \sqrt{\frac{1}{\epsilon}} \tag{14}$$

the ray will in the ionized medium form an angle ψ with the boundary surface, where

$$\mathrm{tg}\,\psi \cong \sqrt{\epsilon}, \qquad (15)^{32}$$

so that $\psi \ll \psi_0$ when $\epsilon \ll 1$.

For refraction, the direction of the wave ray in the two media will be determined by

$$\sin (90° - \psi_0) = \sqrt{\epsilon} \cdot \sin (90° - \psi). \tag{16}$$

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[30] *P. R. W.*, p. 32, Fig. III, 9.

[31] With regard to the limit of the earth's atmosphere we may refer to A. Veronnet, "Constitution et Evolution de l'Univers," p. 153 (Paris, 1927).

[32] *P. R. W.*, p. 18, Table I.

In both cases there is thus a pronounced tendency to transfer the electromagnetic energy from the non-conductive medium to the ionized medium, and further, for the direction of the wave ray in the latter to become parallel to the boundary surface. And this tendency must exist even if the transition from the one medium to the other may not be quite sharp but changes gradually.

The well known far-reaching ability of the short waves may thus be satisfactorily explained by means of the propagation here discussed, but this manner of propagation does not afford an explanation of the long delayed echo signals.

3. The Propagation of Radio Waves in a Space between an Upper Layer, having a Sharply Defined Lower Boundary—and the Surface of the Earth.[33]

This problem has been partly treated in the preceding, but must be considered a little more closely. A ray SP (see Fig. 6) leaving the transmitter S strikes at the point P a part of the upper boundary surface which is not quite concentric with the earth's surface but having such an acclivity that the reflected ray PQ does not come quite down to the earth. If the upper boundary surface taken as a whole is otherwise a spherical shell concentric with the earth's surface, then such a ray may travel on in the manner referred to without ever returning again to the earth. If the intervening space is perfectly non-conductive then the rays may travel round the earth several times since comparatively little attenuation will in general be suffered at each reflection from the upper boundary surface. The number of reflections a during a complete circumscription is

$$a \geqq \pi \sqrt{\frac{R}{2h}} . \tag{1}$$

If the boundary surface is 100 km above the earth and if the radius of the earth is $R = 6400$ km then $a \geqq 17.8 \leqq 18$. The height can hardly be estimated as more than 100 km, since otherwise we should have to reckon with a comparatively strong ionization in the intervening space.

To simplify the case we will consider the losses by reflection at the upper boundary to be negligible. In that case the ray in Fig. 6 will be able to travel very great distances with very small losses; *we are, however, also not receiving any signals.*

If signals are to be received at the earth's surface we may assume as the simplest case that after each reflection from the upper layer the

[33] *P. R. W.*, p. 197–200.

ray returns to the earth, which will be at least 18 times during each circumscription of the earth. The ray may then be either reflected from the earth's surface at an angle of inflection very little less than 90 deg., or it may just graze it. In the first case the amplitude will decrease in the ratio of about 0.8.* In the last case the ray will travel at least about 16 km along the surface of the earth corresponding to that the earth "dips" 5 m inside of the ray. In this manner we get for each reflection an attenuation factor [34] of about $e^{-0.3}$. One circumscription of the earth gives thus an attenuation of respectively $0.8^{18} = 10^{-1.75}$ and $e^{-5.4} = 10^{-2.35}$. After 100 circumscriptions the attenuation would be respectively 10^{-175} and 10^{-235}.

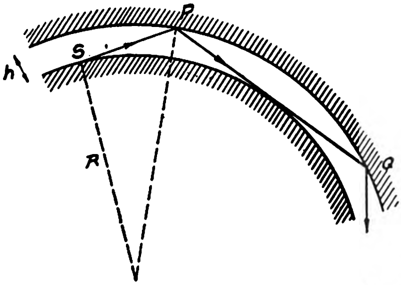

Fig. 6—Schematical representation of a possible path of a ray in the nonconductive space between an upper ionized layer and the surface of the earth.

If we assume that reception is possible within a zone of 40 km at each reflection from, or grazing along, the earth's surface—according to the assumption made above, this width may hardly be estimated at a higher value —then the probability that a signal is received at a given place during one circumscription is equal to $40 \cdot 18/40000 \leqq 1/55$.

This probability is of the right order of magnitude since, if a certain signal is giving a long delayed echo at all, it very often gives two or more;[35] and for 110 circumscriptions we shall then actually have a probability of getting two signals.

On the other hand, the above calculated attenuation is unreasonably great. The attenuation exponent should be reduced to at least

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* *P. R. W.*, p. 132–135, Fig. VIII, 11–18.

[34] *P. R. W.* p. 19, Fig. III, 2.

[35] *Balth. van der Pol*, loc. cit. Of a series of 11 long delayed echoes observed in Oslo the three gave only one echo, six gave two echoes and two gave three echoes. This gives an average of $(21/11) \leqq 2$ echoes for the cases where echoes are obtained at all. The present experimental material is, however, too small to justify a definite conclusion with regard to the average number. In the observations here considered the average of delay was 16 sec.; the longest being 30 sec., the shortest about 3 sec.

1/50 of its value, and consequently also the number of reflections or grazings at the surface of the earth should be reduced to 1/50 of that assumed above in order to get a reasonable attenuation. In that case, however, the probability that a signal will give two successive echoes is reduced also to about 1/50, which is in contradiction to the available experimental material.

Even though the above considerations may not prove the impossibility of obtaining long delayed echoes in the manner last referred to, they show, at all events, that such a probability is extremely small. And there is further another circumstance which is decidedly against such a probability.

The losses due to the reflections at both the upper boundary surface and at the surface of the earth, or by grazing along the earth's surface, will decrease with increasing wavelength. Somewhat longer waves, as for instance 60 or 90 m, should consequently show extraordinarily long ranges under such conditions as are considered here. Two circumstances may, however, possibly counterwork the longer range of the longer waves: Due to the lower frequency the rectilinear radiation would be less pronounced. This, however, is again compensated by a more regular reflection of these longer waves at the uneven surface of the earth. The second circumstance is the attenuation caused by the presence of ions or electrons in the intervening space hitherto considered perfectly non-conductive. No doubt, such ions are present and they will attenuate the longer waves most. The attenuation constant may namely in this case be determined by[36]

$$\gamma_0 = N \cdot \frac{2\pi e^2}{mc} \cdot \frac{\nu}{\omega^2+\nu^2}, \tag{2}$$

which for the 60-m wave will be at most 4 times, and for the 90-m wave at most 9 times greater than for a 30-m wave. Since all of the other attenuations, under the conditions postulated, are smaller for the 60 and for the 90 wave than for a 30-m wave, then the 60-m wave must have a range of at least 1/4, and the 90-m wave at least 1/9 that of the 30-m wave. If the latter is able to travel round the earth for instance 100 times, then the longer waves should be able to do so at least 25 and 11 times respectively. Any such feature has, as far as the writer is aware, never been observed. In all of the cases where complete circumscription of the earth has been observed the wavelength was between 15 and 25 m.

Presumably therefore we may consider it as an established fact that the long delayed echoes are not obtained in the manner last referred to.

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[36] *P. R. W.*, p. 121 (8a).

Also, presumably, we have now tried every conceivable possibility of explaining the long delayed echoes by means of the propagation of radio waves within the earth's atmosphere.

4. The Reflection of Radio Waves From—or Propagation Along—Bands of Ions out in Space.

Since, according to the preceding, the long delayed echoes cannot arise either by the propagation of radio waves within the earth's atmosphere, or by the waves travelling outside the latter in a medium so strongly ionized that the group velocity approaches zero, they must be due to the fact that the waves have travelled very great distances outside the earth's atmosphere. I shall not enter too deeply into the astrophysical problems connected with the present case, but shall only consider some of the relevant wave propagation problems.

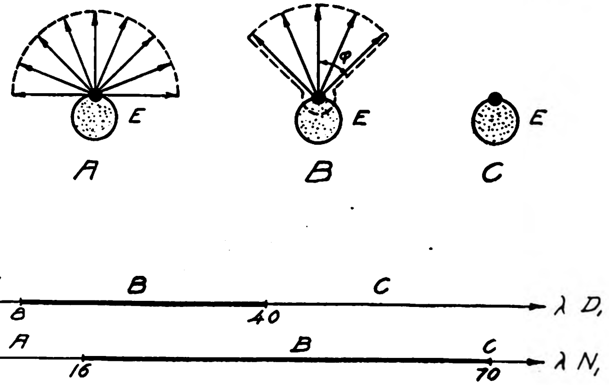

Fig. 7—All wavelengths within the interval *C* (D_1 represents the conditions at noon, N_1 the conditions at midnight; about 40 deg. northern latitude) cannot penetrate out to space but are refracted or reflected back to the earth's surface. All waves within the interval *A* penetrate the atmosphere and go out into space. Within the interval *B* it depends upon the angle of ascent of the wave whether the waves leave the atmosphere or are reflected back again to the earth's surface.

The first problem in that respect is: which radio waves are able to penetrate the earth's atmosphere and proceed out into space? This question is thoroughly treated elsewhere,[37] and according to what is set forth there we assume (see Fig. 7) that all waves shorter than about 8 m will always penetrate out into space with comparatively small attenuation. At midnight this will be the case for all waves up to about 16 m. All waves longer than 40 m are completely refracted or reflected back to the earth at noon, and at midnight all waves longer than about 70 m. The given limits are, however, not fixed values but vary according to the varying state of ionization of the upper atmosphere. All experimental data indicate, however, that the figures given are reasonable mean values.

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[37] *P. R. W.*, chap. XI, especially section (i), p. 214–218. See also "Radiofoniens Aarbog 1929;" p. 16–20 (Copenhagen, October, 1928.)

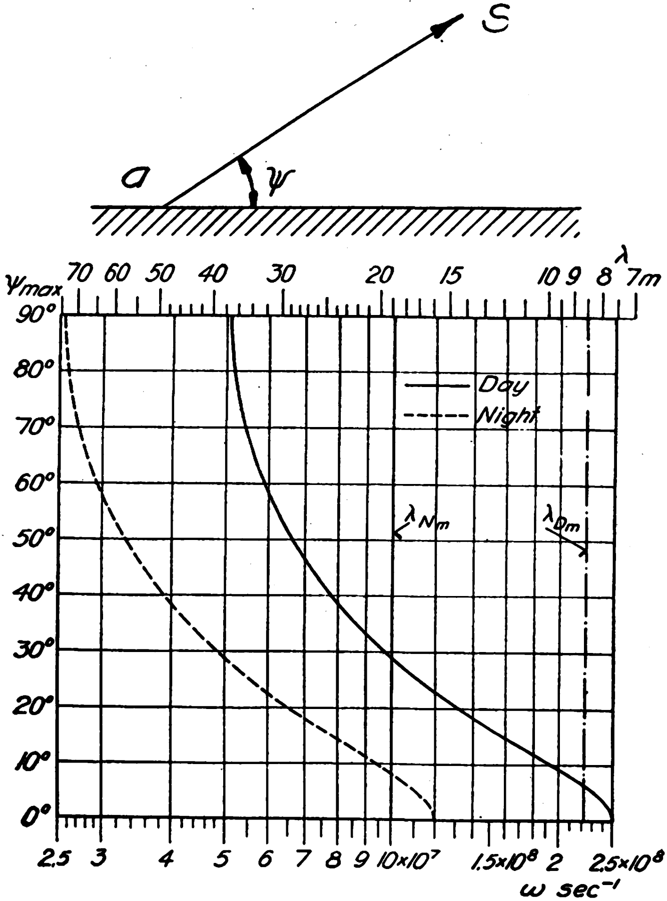

Fig. 8—The earth angle Ψ_{max} as function of the wavelength λ. The full drawn curve represents the conditions at noon, the dotted curve at midnight.

For waves between 8 and 40 m at noon it will consequently depend upon their angle of ascent ψ (see Fig. 8) whether they penetrate out into space or return to the earth. The greatest angle of ascent at which a wave does not penetrate out into space is called ψ_{max}, and the dependency of this angle upon the wavelength at noon is shown in the full drawn curve in Fig. 8, while the dotted curve shows the same relation at midnight. These curves also depend of course upon the state of ionization in the upper atmosphere, but represent reasonable mean values.

According to the above, the long delayed echoes can arise only with waves shorter than 70 m. Waves longer than 70 m can generally neither penetrate out into space nor from the outside penetrate the upper ionized layer and come down to the earth.

The relations for these long waves are indicated in part *C*, Fig. 7, where *E* indicates the earth and the black point the transmitter. No wave rays go from the transmitter out to space.

The relations for very short waves are indicated at *A*. Here emission occurs to the whole of the hemisphere which has for lower boundary the tangential plane to the surface of the earth at the point of transmission.

Of most interest are the relations within the active short-wave interval *B*. Here the emission to space occurs within a cone having the apex angle

$$\phi = 90° - \psi_{\max}. \tag{1}$$

If the transmitter radiates with equal strength in all directions, then the ratio between that fraction η which penetrates out into space and the total radiation within the whole hemisphere will be determined by

$$\eta_1 = 1 - \cos\phi = 2\sin^2\frac{\phi}{2}. \tag{2}$$

If the radiation occurs from a vertical, linear aerial, the length of which is small in comparison to the wavelength, then the radiation intensity is proportional to $\sin^2\phi$ and we then get[38]

$$\eta_2 = 1 + \frac{1}{8}\cos 3\phi - \frac{9}{8}\cos\phi. \tag{3}$$

In Fig. 9 are shown the values of η_1 and η_2 as functions of the wavelength λ. Corresponding values of λ and ψ_{max} are taken from Fig. 8 and then ϕ is found by means of equation (1).

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[38] Beside the direct radiation to space calculated by means of (2) and (3), where the wave rays out in space proceed in nearly the same direction as the one at which they left the transmitter, a more or less regular radiation will take place from—and in a direction nearly parallel to—the upper surface of the ionized layer. This radiation is mainly due to those rays which leave the transmitter at an earth angle in the vicinity of $\Psi_{\max}$. Rays may be emitted from any point of the upper surface of the ionized layer on wavelengths within the interval *B*, Fig. 7, but the intensity of such rays is generally rather small.

From Fig. 9 it appears that at wavelengths appreciably more than 30 m η_1 and η_2 decrease to very small values. Above 30 m the radiation out to space will consequently be comparatively very small, and from the figure it appears at all events that the probability that waves may penetrate out into space in a favorable direction decreases rapidly at increasing wavelength for waves from about 30 m and upwards.

On the other hand, we have shown above that for total reflection at normal incidence of a 30-m wave a density of electrons of about $N_0 = 1.2 \cdot 10^6$ per cm^3 is necessary. At other wavelengths the necessary density of electrons is determined with an accuracy sufficient in this connection by

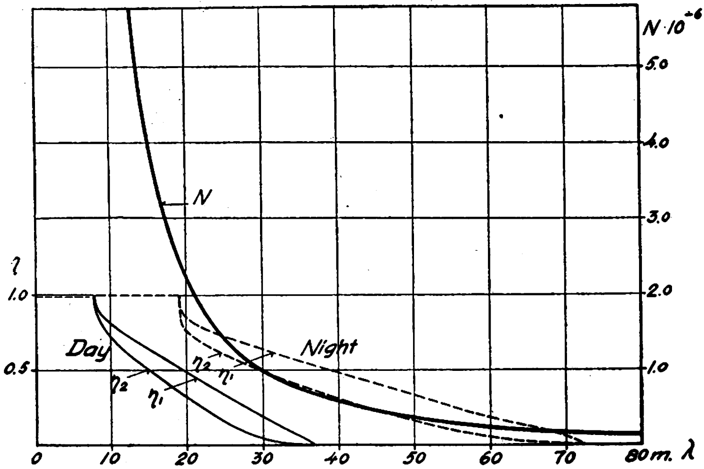

Fig. 9—The values of the constants η_1 and η_2 at day and at night, calculated on the basis of Fig. 8. The N-curve shows the values of N calculated by means of equation (4).

$$N = N_0\left(\frac{30}{\lambda}\right)^2 \cong 1.2 \cdot 10^6\left(\frac{30}{\lambda}\right)^2. \tag{4}$$

The N curve in Fig. 9 shows the values of the density of electrons determined in this manner, and we see that these values increase rapidly with decreasing wavelengths. The probability of finding a band of ions having a density of electrons equal to or greater than N is of course smaller the greater the value of N.

From the preceding it appears that there must be a certain wavelength for which the probability of getting long delayed echoes is the greatest, and from Fig. 9 we may draw the conclusion that this most favorable wavelength must be about 30 m; but the possibility exists of course that echoes may be observed on somewhat shorter as well as on somewhat longer waves; but to the best of our knowledge, long delayed echoes so far have been observed only on waves of about 30 m.

If we were to judge solely from Fig. 9 we would be apt to assume that a somewhat longer wave, for instance about 40 m, would be the most favorable for the observation of echoes. By calculating the η_1 and η_2 curves we have, however, not considered the loss suffered by the waves by their passage through the ionized part of the upper atmosphere, and since this loss increases with increasing wavelength the said curves for the effective radiation out to space should decrease somewhat more rapidly than shown in Fig. 9.

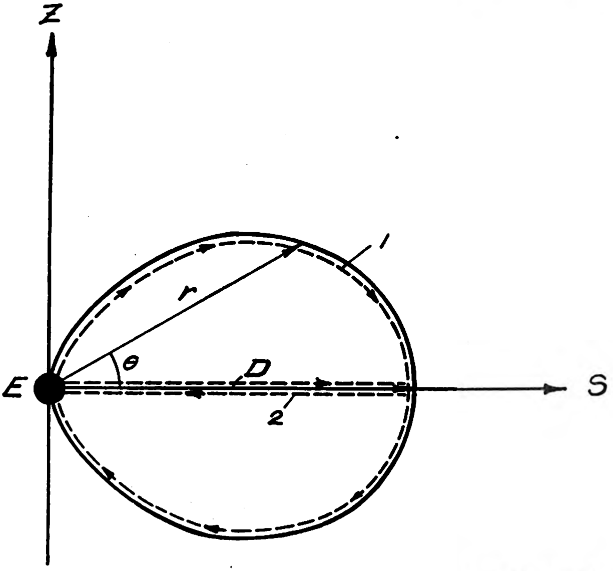

Fig. 10—The full drawn curve indicates the inner limits of the paths of charged particles according to equation (5). *E* represents the earth with its magnetical axis along the *Z*-axis. The dotted curves 1 and 2 indicate the two echo tracks possible, the direction *S* pointing toward the sun.

Another circumstance of some importance in this connection may be mentioned. All of the preceding considerations are based upon a state of ionization of the atmosphere which is assumed by the writer in consequence of the experimental material at hand with regard to the propagation of radio waves in general. Considering this very comprehensive material there is some reason to believe that the ionization distribution assumed as a whole is approximately correct from the earth's surface up to the altitude of maximum ionization, i.e., up to about 130–150 km. As to the density of electrons and the conductivity above that height where the ionization is maximum, our ordinary terrestrial experience with propagation of radio waves is unable to give any information. The density of electrons at greater heights may consequently very well exceed somewhat the values indicated in *P.R.W.* Hitherto it has been possible only to judge about the state of ionization in these regions above the altitude of maximum ionization by means of multiple echoes of short delays[39] and by means of the terrestrial magnetical conditions.[40] The investigation of the ionization of these very high layers may possibly in the future profit by the experiences gained from the long delayed echoes.

With regard to the character and the location of the bands of ionization which cause the echoes, we may just mention that presumably they may be divided into two groups, namely, those along which the waves are propagated, and those from which the waves are reflected.

Of the first kind are presumably those bands of electrons which according to Störmer are due to the invasion of electrons into the magnetic field of the earth; see for instance Fig. 10. As long as it is a question of relatively few electrons only, the classical work of Störmer offers the necessary information with regard to the form of the tracks possible, and to the bounding of those spaces to which such tracks are limited. Above we have shown, however, that in those bands or layers which may be of importance for the phenomena here considered of the propagation of radio waves, the density of electrons is so great that the electrostatic forces will prevent the development of pure electron bands or of dense bands consisting exclusively of ions of one sign. In the ionized bands which are effective echo reflectors the positive and the negative space charges must be of nearly equal value, *i.e.*, the resulting space charge must be comparatively small. Such a band of ions will also have a tendency to follow the tracks calculated by Störmer, and the shape of the track and the velocity will in the main be determined by the heavy ions.

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[39] *P. R. W.*, p. 209–212.

[40] The bearing of the magnetic evidence on the value of the total ionization of the atmosphere is discussed in a valuable paper by S. Chapman just published ("On the Theory of the Solar Diurnal Variation of the Earth's Magnetism," *Proc.* Roy. Soc. (A) **122**, 369–386, 1929.) This author comes to the conclusion that the total ionization required from a magnetical point of view is of the same order of magnitude as that assumed by the writer in *P. R. W.*, the magnetic evidence indicating a somewhat higher value. Chapman adheres to the "dynamo" theory eventually combined with the "drift current theory" which he suggests in the above mentioned paper. Upon numerical considerations he concludes that the "diamagnetic" theory of Ross Gunn (*Phys. Rev.*, **32**, 133–141,1928) cannot be right. It may be added that N. Bohr ("Studier over Metallernes Elektronteori," Chap. IV, Copenhagen, 1911) has proved that such a diamagnetic effect does not exist. The remarks of R. Gunn (loc. cit., p. 136) do not meet the main point of Bohr's arguments and do not invalidate his general conclusions.]

In Fig. 10 we have as an example shown the boundaries of the tracks of the corpuscular rays in a particular case for which the equation of the bounding curve is [41]

$$r=\frac{\cos^2\theta}{1+\sqrt{1+\cos^3\theta}}\cdot\sqrt{\frac{\frac{M}{cm}}{\frac{}{e}\cdot U}} \qquad (5)$$

where r is the radius vector, $(\pi/2-\theta)$ its angle with the magnetie axis Z of the earth. $M=8.6\cdot10^{25}$ = the magnetical moment of the earth, $c=3\cdot10^{10}$ cm sec.$^{-1}$ = the velocity of light in empty space, m and e the mass and the charge of the ion [esu].

According to Störmer the bounding surfaces which may give long delayed echoes constitute the boundary surfaces between those parts of space into which the ions in question may penetrate at a given velocity, and those parts of space into which these ions cannot come. Such boundary surfaces may be efficient as guiding planes for a propagation of radio waves, and such a propagation is indicated by the dotted curve 1 in Fig. 10.

For such a propagation along a band of electrons or ions, which connects the northern and southern polar areas, and which has very great radii of curvature, only a comparatively small density of electrons is demanded and, according to Table IV, the propagation may occur with comparatively small losses.

The other manner of obtaining echoes is indicated by track 2 in Fig. 10. Here the wave ray is reflected from the bounding surface and returns to the surface of the earth. This manner of propagation may occur with only very little attenuation, but demands in the bounding surface the density of electrons determined by (4).

These bands of ions, formed by the magnetic field of the earth, presumably play an important rôle in the production of wireless echoes up to 30 sec. and possibly even up to 60 sec. as was originally suggested by Störmer[42, 43].

But there is also the possibility that outside that space in which the magnetic field of the earth exerts its influence in this manner, bands of ions having sufficient density of electrons may attain such forms that they act as reflectors which after one or more reflections return the radio waves to the earth. If a wave ray is to return to the earth with sufficient intensity after a single reflection from a very distant ionization band the center of the curvature of that band must necessarily be located at or near the earth. Such curvatures may partly be due directly to the influence of the earth's magnetic field and, outside that space where the latter appreciably influences the shape of the ionization bands, partly to the electric field from charges directly on the earth and particularly from such charges which may be "arrested" by the magnetic field of the earth. Such bent ionization tracks are outlined in Fig. 11.

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[41] C. Störmer, loc. cit., p. 129.

[42] C. Störmer, *Nature,* **122**, 681, 1928; *C. R.* **187**, 811, 1928.

[43] For this kind of echo the probability of obtaining good echo signals will, as shown by Störmer (*Nature,* **123**, 16, 1929), depend upon the angle between the magnetic axis of the earth and the direction to the sun, being greatest when these two directions are at right angles to each other. [The latest evidence seems to be in agreement with Störmer's prediction.]

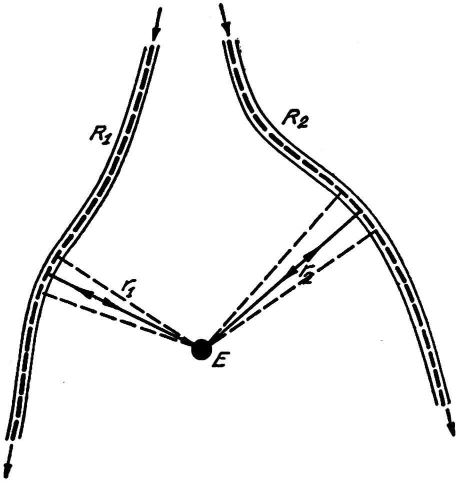

Fig. 11—Two curved ionization bands acting as concave mirrors at the places indicated. *E* represents the earth.

Long delayed echoes may further be obtained in the manner indicated in Fig. 12, where R_1 and R_2 are two bands of ions extending from the sun and which at *a* and *c* reflect a wave ray coming from the earth *E*. Between *a* and *c* the wave ray traverses the curved path *abc*, the curved form of which is due to increasing density of electrons in the direction toward the sun.

Since, generally, the bands of ions are not perfectly electrically neutral they will mutually act upon one another, thus forming more or less curved bands. In this case the center of curvature will, however, in general not be located near the earth and consequently the bent part of the band will not be efficient in producing echoes. If, however, the radiation from the earth is emitted at a comparatively great space angle an exceedingly large space within our solar system will be "searched" and consequently the probability of finding a favorable constellation may be not altogether vanishingly small, although such echo phenomena presumably are rather seldom.

In consequence of the above considerations the writer therefore anticipates that in the future, as the study of the echoes is carried on more systematically and on a more extensive scale there will occasionally be reported echoes of very long delays, possibly up to 10 to 15 min. or even more.[44]

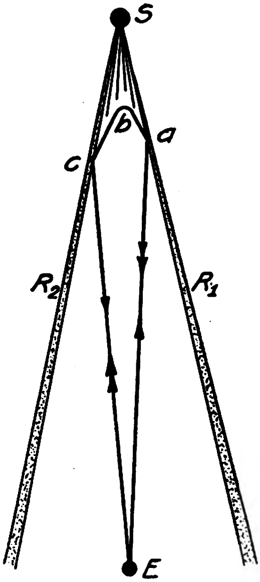

Fig. 12—Schematical representation of a possible manner of obtaining echoes after two reflections (at *a* and *c*), and of a curved path *abc* of a wave ray between the two points of reflection. *E* represents the earth.

From the considerations set forth in this paper, it presumably appears that:

(1) Echoes delayed more than 10 sec. cannot be due to the propa-

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[44] This prediction, set forth some time ago by the writer ("Radiofoniens Aarbog 1929," p. 22–24, October 1928—the above Figs. 8, 11, and 12 are taken from this paper) has received an unexpectedly quick confirmation, since in a letter dated February 2, 1929, Jörgen Hals of Oslo communicates that he has observed echoes up to 4 min. 20 sec. corresponding to a path-length of 78,000,000 km. This observation, if correct, confirms the above considerations, namely, that the long delayed echoes are not due to propagation entirely within the earth's atmosphere, and further that echoes may occur with so long delay that they must be due to ionization bands located outside that space in which the magnetic field of the earth directly exerts its influence. [Another echo, having a retardation of 3 min. 15 sec. is reported by Mr. Hals, February 14, 1929.]

gation of radio waves entirely within the atmosphere of the earth, nor to a propagation of the waves outside this in a medium being so densely crowded with electrons that the group velocity decreases to such small values that the distance travelled will be comparatively short.

(2) Echoes delayed up to 30(possibly to 60) sec. are probably due to propagation along—or reflections from—"Störmer bands" of electrons within the magnetic field of the earth.

(3) Occasionally echoes may be obtained with such great delay that those bands of ions to which the echoes are due must be located at such great distances from the earth (more, for example, than 40.000.000 km) that they are outside the space in which the magnetic field of the earth exerts any appreciable direct influence.

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AN ECHO INTERFERENCE METHOD FOR THE STUDY OF RADIO WAVE PATHS*

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Summary—*An experimental determination of the rate of change of radio-frequency phase of the separate downcoming echoes has been carried out. The small power crystal-controlled oscillator circuit of the 4,435-kc transmitter operates continuously, the high-power pulse transmission being produced by modulating the power-amplifier circuits. A receiver was operated very near the transmitter, having rapid recovery from the paralyzing effect of the pulse ground wave and sufficient sensitivity to receive the echoes with good amplitude, and having a very slight coupling to the crystal-oscillator circuit such that the crystal "pick-up" was comparable to the echoes received. The echoes showed their changing radio-frequency phase by alternately adding to and subtracting from the constant crystal oscillator pick-up. This "interferometer" is naturally sensitive to small changes in the optical path of the waves. The phase changes are regular, but the time of one 360 deg. phase change on 4,435 kc varies from 1 sec. to 60 sec., or possibly longer during the day and evening, and at times changes between these limits in as short a time as 15 min. When multiple echoes are present, the second and third echoes phase in and out more rapidly than the first echo, but not by an even factor.*

IT has been obvious from the time of its inception that the echo-method possesses singular advantages for the study of radio wave propagation, and particularly for the study of the ionized Kennelly-Heaviside layer, by reason of the fact that the waves reaching the receiver by different paths are separated, instead of being received simultaneously, provided, of course, that the transmitted pulses are of sufficiently short duration. The complex echo pattern received on 4,435 kc, especially during the evening and night hours,[1] has emphasized the importance, at least in this region of frequencies, of such an analysis of the energy received over the various paths. By suitable arrangements the polarization, phase, amplitude fading, and other characteristics of the waves received over each separate path can be studied.

We have recently carried out an experiment which determines the rate of change of radio-frequency phase of the e.m.f.'s in the receiver due to the separate downcoming echoes. This has been done by allowing the separate echoes to combine at the receiver with a small (radio-frequency) "pick-up" from the transmitter crystal. At present the experiment has been carried only far enough to verify the possibility of studying radio wave paths by such a method, but the results obtained seem interesting enough to justify a brief report. The small power, crystal-controlled oscillator circuit of the 4,435-kc transmitter operates continuously, the high power pulse transmission being produced by modulating the power-amplifier circuits with an unbalanced multivibrator.[2] A receiver was operated very near to the transmitter, the receiver having rapid recovery from the paralyzing effect of the pulse ground wave and sufficient sensitivity to receive the echoes with good amplitude and having a coupling to the crystal-oscillator circuit such that the crystal pick-up was comparable to the echoes received. The echoes showed their changing phase by alternately adding to and subtracting from the constant crystal-oscillator pick-up.

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[1] L. R. Hafstad and M. A. Tuve, Proc. I. R. E., **17**, 1513; September, 1929.

Before the experiment was performed it could not be predicted with certainty that the transmitter crystal would provide a suitably constant phase reference for comparison with the echoes, particularly because of the fact that its phase and even its frequency might be shifted slightly by the pulse modulation of the power amplifiers excited by the crystal circuit. We have observed, however, that with great regularity the echoes phase in and out with respect to the crystal pick-up. A constant phase shift of the crystal with each pulse, of course, would not be detected, nor would it affect the interpretation of the results of the experiment as measuring the changes in optical path for each echo. It is possible that the phase of the crystal between pulses may be variable with respect to the phase during emission. This point remains to be studied, but if so the observations here reported do not correctly measure phase changes due to changes in optical path. Nevertheless, since the separate multiple echoes from a given set of pulses phase in and out at different rates, and as it is very unlikely that the phase of the crystal is disturbed in the intervals between multiple echoes, at least part of the observed phasing must be due to changes in optical path. It is difficult to imagine that the crystal phase or frequency could have been changing *progressively* in such a way as to cause the regular phasing which was observed. In addition to possible sudden phase shifts due to the modulation, another source of error to be considered is that involved in the change of the crystal frequency with temperature. One may question whether the electrical phase of the crystal alters appreciably from its correct value as a reference phase, due to changing temperature, in the interval between a pulse and its echo. The crystal frequency changes with temperature about one part in 40,000 per deg. C. To change the electrical phase by even 1/1000 radian from its correct value, in the 0.002-sec. interval between a pulse and its echo, would require the temperature to change at the rate of about 1 deg. C per sec. Actually the temperature of the crystal does not alter, at most, by more than 3 deg. to 5 deg. C during several hours. A second point to be considered is whether a slowly changing frequency, due to changing temperature of the crystal, might produce phasing of the echo with respect to the crystal, for reasons similar to those underlying the wavelength change method of Appleton and Barnett,[3] i.e., phasing of echo with respect to crystal due to the varying frequency causing changes in optical path and changes in the exact number of crystal cycles taking place between pulse and echo. However, a phase change of 360 deg. observed to occur on the average in perhaps 10 to 15 sec., means a change in optical path or number of cycles between pulse and echo of approximately one part in 3,500. Since a temperature change as large as 1 deg. C causes a frequency change of only one part in 40,000, it is clear that only a negligibly small rate of phasing of the echoes can be caused by temperature (frequency) change of the crystal, unless the frequency happened to pass through a possible critical value at which very large path changes might be caused by very slight changes in frequency.

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[2] M. A. Tuve and O. Dahl, PROC. I. R. E., **16**, 794–798; June, 1928.

The receiver was located about 300 feet from the transmitter, with the two antennas approximately at right angles. The apparatus used was entirely similar to that previously used here for echo observations,[4] with the exception that grid bias detection was substituted for grid leaks and condensers, and moving-picture film was used for recording, giving a record of any desired time length. Continuous records of 15 sec. or longer were found most suitable. Visual counts giving the 360-deg. phasing time were found to be a valuable supplement to the records, especially for the phasing time of the first echo. In fact, with our present arrangements suitable photographic records of phasing times over a period of several hours would require large quantities of film. Low time constants throughout the receiver, giving very rapid recovery, were of obvious importance so near the transmitter, but as careful elimination of condensers in the receiver and the use of grid bias detection instead of grid leaks had been resorted to when first using multivibrator modulation, no trouble was experienced in this respect. Using super-speed motion-picture film an ordinary 6-volt tungsten lamp gave a satisfactory trace with an optical lever of about

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[3] See *Proc.* Phys. Soc., London, **41**, 43; December, 1928, and references there quoted.

[4] G. Breit and M. A. Tuve, *Phys. Rev.*, **28**, 554–575; 1926.

50 cm. Timing marks were obtained using an auxiliary spot of light interrupted by a two-sectored disk mounted on a 25-cycle synchronous motor. Film speeds of about 25 and 50 cm per second were used, the latter giving much the better records.

A typical section of the record obtained is shown in Fig. 1. The oscillograph is deflected by the radio-frequency pick-up from the transmitter crystal-oscillator circuit to about half the saturation value for the power amplifier of the receiver. Following the ground peak, which shows a moderate overthrow due to underdamping of the oscillograph element, three echoes are present in this specimen of the record, with retardations equivalent to multiples of about 250 km layer height, as usual. The first is "in phase" with the crystal, deflecting toward saturation, and the second and third are "out of phase" with the crystal, deflecting toward zero. The echo is strictly in phase with the crystal only for an instant during each 360-deg. phase change, of course. The terms "in phase" and "out of phase" are used here somewhat inexactly to refer to the periods during which the resultant of the crystal vector and the echo vector is greater or less than the crystal vector alone. When the ray path is increasing, for example, the echo may be represented by a vector which is rotating, and possibly also changing in amplitude, with respect to the constant crystal vector. During our observations the echo amplitude was always less than that from the crystal. For the simple case of a constant amplitude echo, the deflection due to the echo should be zero (resultant vector equals crystal vector) twice during each 360-deg. phase change of the echo, and due to the difference in amplitudes these points both should lie between 180 deg. and 270 deg. from the point of exact phase, that is, the echo should be "in phase" a longer time than it is "out of phase." Amplitude fading of the echo, possible polarization of the downcoming wave, which may also be changing, antenna directivity, non-linearity of the receiver, and other variables prevent an exact analysis of our present records. However, it seems very probable that the *average* 360-deg. phasing time obtained from the records should closely approximate the actual average time of increase (or decrease) of the optical path of the echo by one wavelength.

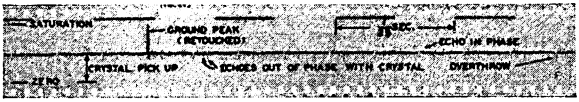

Fig. 1—Typical record showing echoes in phase and out of phase with crystal at 17 h. 55 m., March 3, 1929 (effective heights 250–500–780 km).

Fig. 2 shows a series of sections cut from a single record of about 15 sec. duration, to illustrate the phasing in and out of the first echo. The ground peak has been retouched to make reproduction possible. The time for each section is given in sec. Due allowance must be made for the overthrow of the oscillograph with large amplitudes, due to underdamping. From 0.0 to 0.94 sec. the first echo is out of phase with the crystal; it is in phase from 1.26 to 4.45 sec., out of phase from 5.07 to 6.63 sec. in phase from 7.25 to 8.77 sec., and out of phase from 10.5 to 12.4 sec.

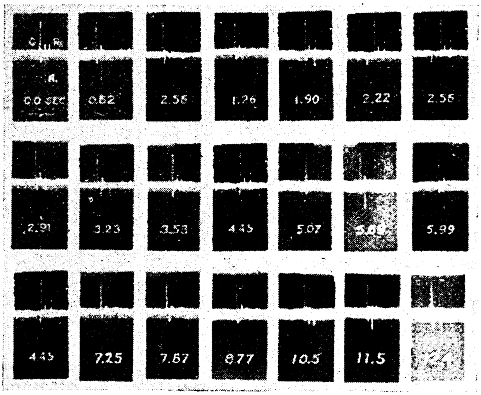

Fig. 2—Successive sections of typical record showing phasing of first echo with more rapid phasing of second echo also evident; time of each section given in seconds. Recorded at 20 h. 19 m., March 3, 1929 (effective heights 250–480 km).

The phasing of the echo was always quite regular, alternately adding to and subtracting from the deflection due to crystal pick-up, but the time for one complete 360-deg. phase change (change of one wavelength in optical path) varied greatly during any extended period, as illustrated by Table I for Sunday, March 3, 1929.

TABLE I
Phasing Times for First Echo

Time P. M.		Average time for 360-deg. phase change, first echo	Time P. M.		Average time for 360-deg. phase change, first echo
h.	m.	sec.	h.	m.	sec.
3	52	15	7	38	5.4
4	18	16	7	51	4.0
4	24	4	8	15	3.0
4	37	7.5	8	23	20
4	52	11.5	8	29	5.0
5	16	3.3	8	44	22
5	21	3.5	8	51	30
6	37	18	9	02	4.3
6	47	50	9	15	5.4
6	56	7.5	9	24	20
7	09	4.8	9	35	4.
7	19	5.4	9	53	30

During the hours covered by Table I the layer height, measured by the retardation of the first echo, changed from 240 km to 300 km approximately. Due to low film speed the accuracy of these figures is not better than about ±12 km. A change of 60 km in effective height would mean an average path increase of about one wavelength (360-deg. phase change) in 12 sec. Averaging the figures given in Table I is obviously impossible, but at least the phasing time would seem to be of the right order of magnitude.

The phasing time for the first echo has been observed to be variable and usually of the order of 10 sec. during midday hours as well. It is perhaps not to be expected that the wave path should be constant at any time, and of course, with the present method of recording, slow fluctuations in the optical path would show phase changes similar to those due to a steady increase or decrease.

We have only a comparatively small number of records, made during the evening when marked multiple echoes are present, which give suitable measurements of the phasing times of the second and higher order echoes. Those we have, however, fail to show any obvious relation between the phasing times of the various echoes. The second and third echoes phase faster than the first echo, but not by even factors of 2 or 3 as might be expected if they were due to multiple reflection over the same ray path. The retardation times for these records were usually multiples to within the errors of measurement, however.

The times for successive 360-deg. phase changes, measured on the film by the distance between successive passages of the echo amplitude through the "shifted zero" (resultant vector equals crystal vector, that is, no echo peak visible) necessarily cannot be accurately determined, but seem to vary much more erratically for the second and third echo than for the first. Amplitude fading should affect the individual measurements, but the average measured phasing time should not be affected by fading unless fortuitously the phasing time is related to the fading time in a particular way. We have observed that at different times for the various echoes the fading time may be much longer or considerably shorter than the phasing time, with no apparent relation between the two.

A fundamental source of trouble in measuring phasing times for multiple echoes accurately was the fact that at times when multiplicity appeared the reflections were often doublets which were too close to be completely resolved. The two echoes comprising a doublet phased independently, giving rise to indefinite echo amplitude "zeros" (resultant vector equals crystal vector). A fairly well resolved doublet, with the components phasing separately through "zero," is shown in Fig. 3. At 0.0 sec. both components are in phase with the crystal, at 0.49 sec. the first component is in phase and the second component is out of phase, at 2.07 sec. the reverse is true, and at 5.41 sec. both components are out of phase. Long records, made with a higher film-speed than 50 cm per second, will be necessary before quantitative statements can be made concerning the relative phasing times of multiple echoes.

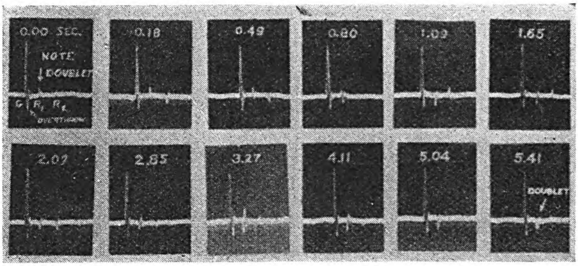

Fig. 3—Successive sections of record showing phasing of doublet echo through the "shifted zero." Recorded at 22 h. 40 m., March 3, 1929.

A brief attempt was made to determine the direction or sense of the phase changes of the first echo, in order to distinguish between increase and decrease of optical path (ionized layer rising or falling) by obtaining records alternately on two antennas, using a relay switch. The results obtained were too indefinite to be of significance, especially as the antenna constants were undefined, but it may be of interest to point out that quite definite information on this point could be obtained by introducing a known periodic phase change in the pick-up from the crystal circuit.

This experiment was performed at the suggestion of our colleague, G. Breit, to whom we are also greatly indebted for discussion and other assistance. The necessary special schedules of signals were very kindly transmitted by the members of the staff of the U.S. Naval Research Laboratory, who have provided the special transmissions for all observations by the Department of Terrestrial Magnetism using the echo method. For this experiment a room was also provided at the Naval Research Laboratory in which our recording apparatus was operated. We are grateful to Captain Oberlin, Dr. Taylor, Dr. Hulburt, and others there for their very cordial cooperation, and to J. A. Fleming of the Department of Terrestrial Magnetism, whose interested support has made the work possible.

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ON THE RELATION BETWEEN LONG-WAVE RECEPTION AND CERTAIN TERRESTRIAL AND SOLAR PHENOMENA*

By
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Summary—*It is shown in the paper that there appears to be an inverse relation between long-wave reception on 75 kc per sec. and the temperatures at the transmitting and receiving stations, when the distance between them is not great. The explanation is offered that the variations in the temperatures and in the signal strength are due to a common source, the changes in the medium between the two places.*

If at all, only a direct relationship has been noticed between field intensity and the barometer reading at the receiving station.

There is found to be good correlation between atmospheric ozone in northwestern Europe and Bangalore reception, and between the two and sunspots. Little correlation is found between these and terrestrial magnetism as represented by the international magnetic character of day numbers. Analysis of the data on the basis of the 27-day period has yielded very satisfactory agreement between reception and sunspots, with a six-day period in reception for which no explanation is found.

The lag of one to two days between reception and sunspots occasionally occurring has been explained on the basis of the emission of high speed corpuscles from the sun. Regarding some of the differences between long- and short-wave transmission, the transmission of short waves is attributed to electron movement in the higher regions of the upper atmosphere and that of long waves to less mobile ions below. It is on this basis that the correlation between reception and ozone may be explained.

The paper is in connection with the signal intensity measurements of Madras (Fort) Radio working on 75 kc per sec., made at the Radio Laboratory of the Indian Institute of Science, Bangalore, between March 1926 and August 1927, a period of eighteen months.

1. Introduction

DURING the last few years, attention has often been drawn to the possible connection between the strength of received radio signals and some of the terrestrial and solar phenomena. In the undoubted influence of the sun on radio wave transmission, interest has been centered on the understanding of the relation of the spots on the sun to the field intensity of distant radio stations. Of the terrestrial phenomena are (a) local temperature and pressure, (b) atmospheric ozone, and (c) the earth's magnetism, with which relation is sought. The success that has attended long distance communication by the use of very high frequencies and some of the opposite effects observed in long and short waves by the same cause have stimulated interest in the study.

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* Dewey decimal classification: R113.5. Original manuscript received by the Institute, January 10, 1929.

In a general way, the problem can be considered to be a statistical one, depending on data accumulated for a considerable period over as much as possible of the radio spectrum, parallel to those on phenomena with which relation is sought. Such complete data do not exist for a short period of even a year. The statistical study has, for some time, receded into the background, due to the fruitful results obtained by the application of optical methods to the study of wave propagation sometimes under special conditions, as in the experiments of Breit and Tuve, Appleton and Hollingworth.

Of the statistical investigations, those of G. W. Pickard[1] and of L. W. Austin[2] are important both for their reliable observations and the fairly long period of time which they cover. Of special interest are the observations of Austin which date back to 1914.

Daily measurements of the received field intensity of Madras (Fort) Radio (VWO), on a test signal transmitted at 0800 I.S.T., have been going on since March 1926 at the Radio Laboratory of the Indian Institute of Science, the frequency of the transmission being 75 kc per sec. Bangalore is about 295 km from Madras, almost directly to the west of it. The method of measurement and some preliminary observations have been described elsewhere.[3]

Although the results in the paper relate to a period of eighteen months, the conclusions drawn can only be tentative in character and cannot partake of the nature of established facts. There have been occasional breaks in the field intensity measurements; in August 1926, the monsoon was responsible for the breakdown of the apparatus for some period, due to the high humidity in the air; traffic pressure during the latter half of December and the early part of January rendered the transmission of the test signals very irregular.

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[1] G. W. Pickard:—
- (a) "Correlation of Radio Reception with Solar Activity and Terrestrial Magnetism," Proc. I. R. E., **15**, 83; February, 1927.
- (b) "The Relation of Radio Reception to Sunspot Position and Area," Proc. I. R. E., **15**, 1004; December, 1927.
- (c) "The Correlation of Radio Reception with Solar Activity and Terrestrial Magnetism, II," Proc. I. R. E., **15**, 749; September, 1927.
- (d) "Some Correlations of Radio Reception with Atmospheric Temperature and Pressure," Proc. I. R. E., **16**, 765; June, 1928.
- (e) "Discussion on Long Distance Receiving Measurements at the Bureau of Standards," Proc. I. R. E., **15**, 539; June, 1927.

[2] L. W. Austin:—
- (a) "Long Wave Radio Measurements at the Bureau of Standards in 1926 with Some Comparisons of Solar Activity and Radio Phenomena," Proc. I. R. E., **15**, 825; October, 1927.
- (b) "Radio Atmospheric Disturbances and Solar Activity," Proc. I. R. E., **15**, 837; October, 1927.
- (c) "On the Influence of Solar Activity on Radio Transmission," Proc. I. R. E., **16**, 166; February, 1928.

[3] K. Sreenivasan:—
- (a) "Madras (Fort) Radio Field Intensity Measurements at Bangalore," *Jour.* Ind. Inst. Sc., **9B**, 37, 1926.
- (b) "Intensity Variations of Madras (Fort) Radio Station," *Jour.* Ind. Inst. Sc., **10B**, 35, 1927.

2. METHODS OF ANALYSIS

The methods of analysis followed in arranging and correlating the data of the various elements are mainly two. (a) Periodic averages of a weekly or monthly type or of both, are compared with respect to the variations therein. (b) With any one of the elements as a reference quantity, the variations of that quantity and of reception on either side of the maximum or minimum days of the former are studied. This method has been extensively used, a very interesting instance being the paper by Chree and Stagg in a study of the 27-day period in terrestrial magnetic phenomena. Further, to get rid of comparatively transient changes and bring the main similarity or dissimilarity between the two quantities under examination, considerable use has been made of the smoothing formula $(a+2b+c)/4$.

3. TEMPERATURE AND SIGNAL STRENGTH

The daily local temperature at Bangalore has been obtained from the Bangalore Meteorological Observatory and relates to 0819 I.S.T., 19 minutes after the signal has been transmitted. Those for Madras for the same time were obtained from the Madras Observatory. That the temperature of the room in which the signal is measured is not necessary will be evident at a later stage.[4]

Practically every experimenter in field intensity measurements has come to the general conclusion that the months October to January form a period of very pronounced changes in the daily intensities of radio stations, especially of long waves, in the band 100 kc per sec. to 12 kc per sec. At moderate distances, as Hollingworth[5] showed, the change may be either a decrease or an increase depending upon the position of the receiving station in the interference pattern produced by the ground and the downcoming rays. In general, at fairly big distances, the variations are an increase, including in this statement the Bangalore observations on Madras, though with only 295 km between them.

It is exactly this period of the year which is characterized by low temperatures, the hotter seasons being characterized by low field intensity. Austin's curves (loc. cit.) for Tuckerton and New Brunswick are very good examples of this.

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[4] C. Chree and Stagg, "Recurrent Phenomena in Terrestrial Magnetism," *Roy. Soc., Phil. Trans.*, 227, 21–62, Aug. 3, 1927.

[5] J. Hollingworth, "Propagation of Radio Waves," *Jour.* I. E. E., **64**, 1926.

To determine if the interrelation between the two is purely restricted to the receiving station temperature or if there is any relation with the temperature at the transmitting station as well, the two temperatures were plotted together with the rather unexpected result that the curves march almost parallel to each other. Evidence of this can be seen in the detailed curves *a* to *j* of Fig. 1, where the daily deviations

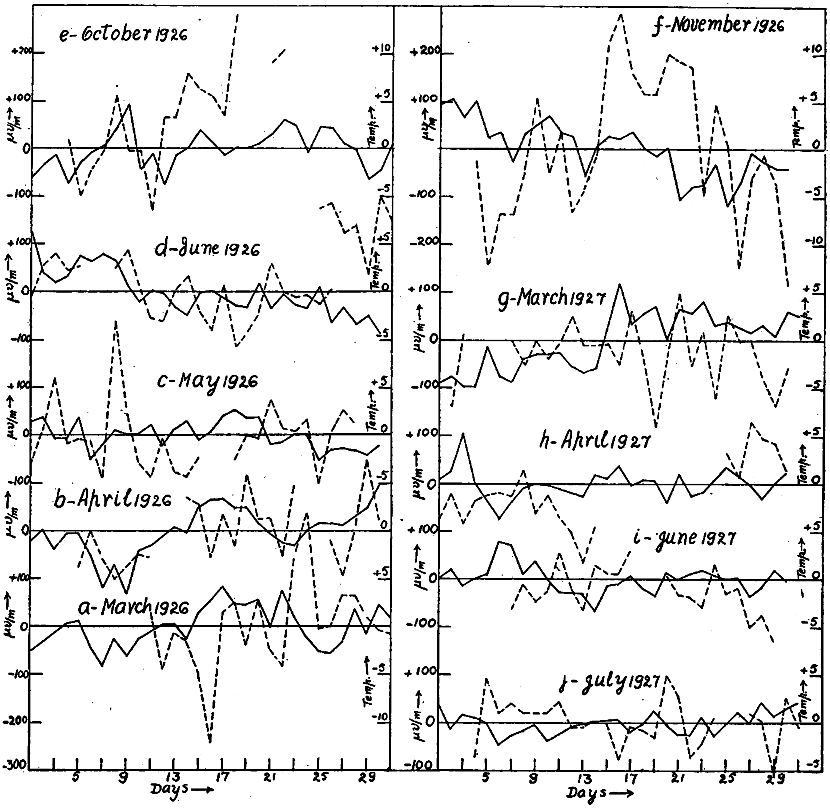

Fig. 1

from monthly averages are plotted. Even in the matter of details, the agreement between the weekly averages plotted in Fig. 2 as the ratio 100/*T* is very close. With monthly averages, the parallelism between the two temperature curves is hardly broken. (Fig. 3)

These curves show a broad inverse relation between temperature (at either station) and field intensity. Of closer similarity there is none, due obviously to the masking influence of other factors of both terrestrial and solar origin. That the inverse relation is not a purely local matter, but is due to changes that are common to the temperature throughout the distance between the stations is the chief point of interest. In view of this, it is legitimate to expect that in the results of Austin too, the transmitting station temperature will have parallel changes with those at the receiving station.

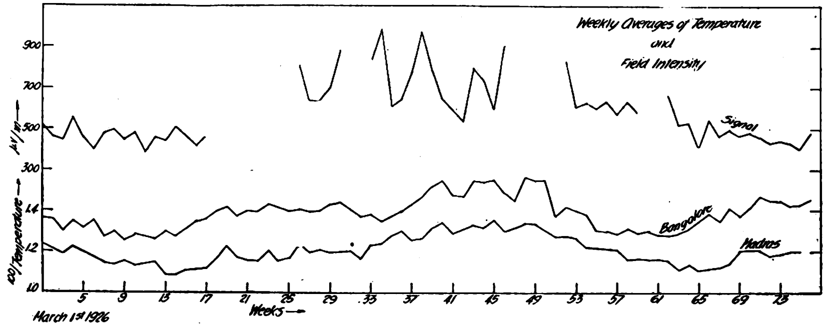

Fig. 2

The intervening medium between the two places undergoes changes which are reflected in those of the temperatures at the two stations, so that the inverse relation between field intensity and temperature is not strictly of a local character but is due to the non-local type of changes occurring between the two stations.

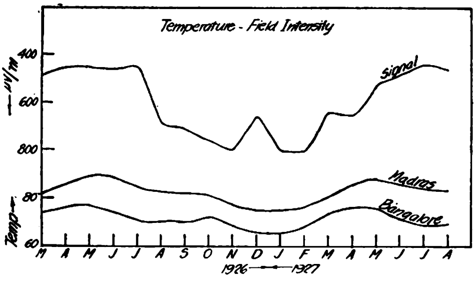

Fig. 3

Of interest are the curves of daily departures from monthly means, although they sometimes give contradictory results (Fig. 1). During some months, such as November 1926, the daily variations of intensity are very big; but they are not accompanied by as pronounced changes in temperature. In this month, Bangalore temperature shows a gradually diminishing tendency; the field strength decreases in the first week, then increases rapidly and as rapidly gets low with a suggestion of a six-day period. During March 1927, however, the temperatures at both places are generally on the increase while signal strength variations, though big, show an increasing and then a decreasing character just opposite to what we had for November, 1926.

Although it may appear contradictory to the inverse relationship, the curves for the months April, June, and October 1926 indicate that the broad variations in both temperature and field intensity are similar; but during the "steady" months, especially in May 1926, when the day-to-day intensity variations are comparatively smaller than in the winter months, the inverse relation is better brought out. April 1927 is another instance of the "steady" months.

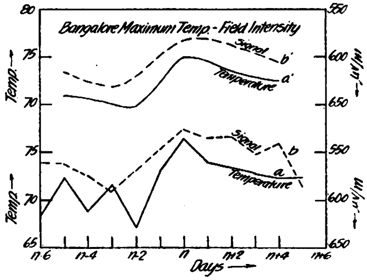

Fig. 4

The second method of study is to select the day in the month on which the temperature has been highest for the whole month. Call this day n; the day previous to it is $n-1$ and the one after it $n+1$ and so on. The temperatures on these days are written under the corresponding number and the average of each is calculated to give the general variation of temperature on either side of day n, for the whole period of 18 months. For these same days of each month, the field strength is also correspondingly written down and the average variation of signal intensity is calculated from these readings. In Fig. 4, the curves a and b refer to temperature and field intensity, respectively. The scale for b increases from top to bottom, in order to indicate better the inverse effect. The curves a' and b' in the same figure obtained by using the smoothing formula $(a+2b+c)/4$ show in a much more pronounced manner that as temperature increases, field intensity decreases and vice versa.

Fig. 5 relates to a similar analysis with minimum temperature day. Here the similarity between the two quantities is very slight, with only a trace of the inverse.

Whether taken over long periods or judged by the second method of average daily variations, there is evidence to conclude that increase or decrease of temperature is generally attended by the opposite variation in signal intensity.

The question naturally arises whether such an inverse effect is evident with long distance observations, say, over more than 1500 km. Signal intensity variations are primarily connected with the state of ionization in the upper regions; the duration of sunlight and the angle of sun's rays at any two places far from each other vary so much all along the path that any similarity in temperature variations, due even to this cause alone, is non existent. So far, in long distance measurements, no results in this matter have been obtained, due certainly to numerous other influences that prevent any useful comparison.

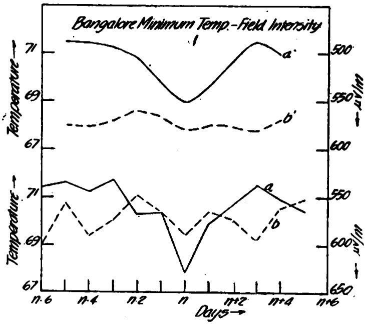

Fig. 5

Temperature and Atmospherics. On the other hand, the close correlation between local temperature and intensity of atmospheric disturbances found by Austin (loc. cit.) is perhaps due to the fact that the latter are bound up with the electric potential gradient of the air in the locality. Proof of this is forthcoming in the observations made at the Central College, Bangalore, by Prof. A. Venkata Row Telang. Low humidity and high temperatures produce high potential gradients of as much as 300 volts per meter or more, even at comparatively low heights from the ground; and atmospheric disturbances are likely to be violent under such conditions. Disturbances due to high potential gradients of a strictly local character are more numerous than those due to others and are distinct from the ones caused by agencies like near or distant thunderstorms.

4. BAROMETRIC PRESSURE AND FIELD INTENSITY

In examining the dependence or, more correctly, the relation between field intensity of Madras and the changes in the barometer readings (supplied by the Bangalore Observatory) at 0819 I.S.T., we notice in the curves for monthly averages (Fig. 6) what appears to be a direct relationship between the two. Between March and July, 1926, the pressure steadily decreases while signal intensity is practically constant. July is also characterized as being the month of lowest barometer reading for the whole year. From July onward, however, there is good measure of similarity between the two curves. The smoothed curves *a′* and *b′* of Fig. 6 give the main variations for the whole period under queston.

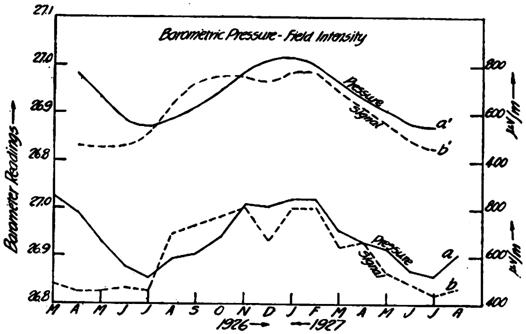

Fig. 6

As with temperature, Figs. 7 and 8 respectively give the average daily variations of pressure and field intensity on either side of the maximum and minimum pressure days of each month for the eighteen months from March, 1926, to August, 1927.

The curves *a* and *b* in the two figures may apparently lead one to think that there is an inverse relation between pressure and field strength. But the curves of monthly averages have distinctly shown a direct connection. That this is more probable will be realized if a forward shift of two days is given to the field intensity curve. This has been done in Fig. 7, giving *a′* and *b′*, and they show the appreciable similarity between pressure and signal intensity. The two-day shift has been incorporated in the smoothed curves *a″* and *b″*.

With the variations of the two quantities on either side of the day of minimum pressure as shown in Fig. 8, there appears to be little agreement. The forward shift too appears to be one day instead of two. However, the interesting fact is that pressure always lags behind reception, indicating that pressure variations are in no way the cause of signal strength changes.

The statement occasionally made that there is no connection between pressure and reception is perhaps due to this uncertain lag, which undoubtedly causes indefiniteness except in cases of prolonged observations. Even then, the relationship is none too marked, nor absolutely definite.

Unlike pressure, there is no lag between temperature and field intensity as we have already seen. The day of lowest temperature is also the day of highest signal strength and vice versa. It is therefore difficult to say whether or not reception changes are caused by temperature variations.

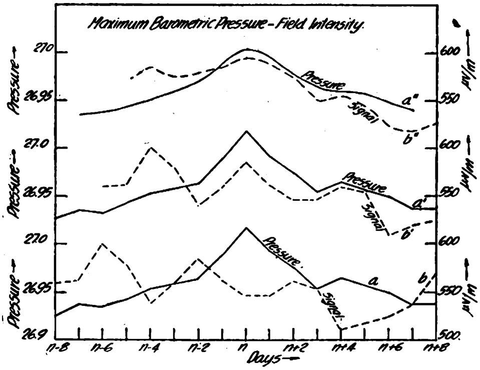

Fig. 7

5. ATMOSPHERIC OZONE AND FIELD STRENGTH

The interest, evident of late, in regard to the presence of ozone in the atmosphere of the earth is shown by the attempts that have been made to understand its relation to a variety of things, such as weather, solar phenomena, and earth's magnetism. The formation of ozone by light in the band spectrum of oxygen from 1300 to 1850 A.U. and the breaking up of ozone into oxygen between 2300 and 2900 A.U.,—the latter figure corresponding to the limit of the sun's ultra violet spectrum[6]—are properties which are likely to have definite relation to the state of ionization of the upper atmosphere.

Apart from the ozone values determined at about half a dozen places in northwestern Europe by Dobson[7] and his associates, there is little information to be obtained. Of the condition of ionization of the gas, the hourly changes in its ionization, the velocity of the ionized molecules in the field of the earth, any changes in the average height at which the ozone layer exists, and other information, nothing is known. Further, there seems to be no method yet devised to determine these. It seems possible that radio wave propagation experiments might be of some help in this direction.

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[6] E. O. Hulburt, "Ionization in the Upper Atmosphere of the Earth," *Phys. Rev.*, 31, 1018, 1928.

[7] The author understands that ozone measurements have recently been started at Kodai Kanal and a few other places in the lower latitudes.

Clayton[8] pointed out the possible existence of some relation between atmospheric ozone and solar phenomena. The properties of the gas in relation to ultra violet radiation from the sun immediately suggests an examination of its possible relation to radio reception. With the figures for ozone values for northwestern Europe kindly supplied by Dr. Dobson, an analysis of the variations of Madras signal strength in relation to the ozone content of the atmosphere was undertaken and the results communicated in a letter to *Nature.*[9]

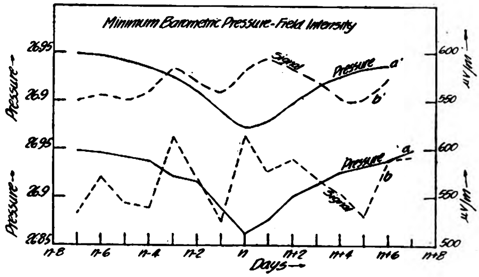

Fig. 8

The period of comparison was limited to the six months from March to August 1927, which was the only interval common to both the measurements. The conclusions arrived at in the paper regarding this matter would have been more reliable if comparison over the whole period were possible, especially with local values of ozone.

The curves *a* and *b*, Fig. 9, represent the variations in the weekly averages of field intensity and ozone value. Except for the ninth and tenth weeks, the curves march almost parallel to each other, particularly from the eleventh week. The reason for the sudden shooting up of signal strength in the ninth week and its equally rapid decrease in the tenth is not understood, as no corresponding pronounced change occurs in any other terrestrial or solar elements. The parallel changes in the two quantities show up much better in the curves *a′* and *b′* (Fig. 10) of smoothed values. If the values for the two abnormal weeks are omitted, the correlation, in the smoothed values, between field intensity and ozone is 0.88 ± 0.023, representing a very satisfactory measure of agreement, in view of the nature of the phenomena.

[8] H. H. Clayton, "Atmospheric Ozone and Solar Variability," *Nature*, 120, 153, 1927.

[9] K. Sreenivasan, "Long Wave Radio Reception and Atmospheric Ozone," *Nature*, 122, 613, 725, 1928.

The inference naturally arises that there is a direct relation between the ozone present in the air and long-wave intensity even when averages over short periods like a week are taken. From the ionization theory of propagation, as far as it is understood, it appears that an inverse relation might be anticipated between ozone and short-wave reception. This is likely, as very high frequency waves in traveling into much higher heights than that at which the ozone layer is estimated to exist get attenuated in penetrating through it.

Long-wave intensity is proportional to the conductivity of the conducting layer. The curves therefore indicate that the variation of the ozone content is proportional to that of conductivity. From this, it appears that the ionization of the ozone in the atmosphere is proportional to the ozone content therein.

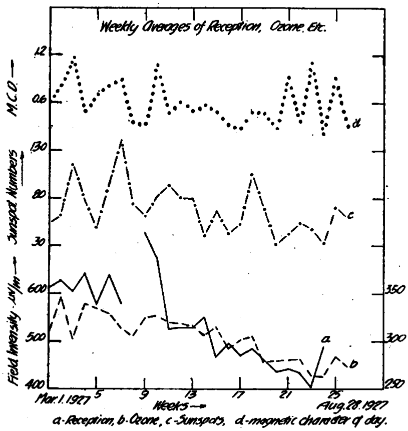

Fig. 9

It is appropriate that, during the period of similarity between field intensity and ozone, a comparison should be made with the variations of the magnetic elements. Curve *c*, Fig. 9, gives the weekly averages of Wolfer's sunspot numbers, while *d* shows those of the international mean magnetic character of day. But for the peak of the eighteenth week, the sunspot curve shows the same general tendency to decrease as the ozone and intensity curves do. After the eleventh, the peaks and troughs in the ozone and sunspot curves occur together, though the ratio of amplitudes in the variations is not constant. The corresponding smoothed curve is *c′*. A further examination into this question cannot fail to yield interesting results.

It is explained that variations in sunspots generally mean changes in the high velocity (charged or neutral) corpuscles and the electromagnetic radiations streaming out of the sun. If a correspondence between ozone and sunspots is quite definitely established, it will facilitate a better understanding of the relation between reception and sunspots. On the other hand, there seems to be no tangible relation between field intensity or ozone to terrestrial magnetism, as represented by the international mean magnetic character of day numbers, as curves *d* in Figs. 9 and 10 show. Neither in general variation nor in details is there any direct or inverse relation that can be definitely traced. This is the more remarkable as one would think that terrestrial magnetic changes would be expected to be accompanied by fluctuations in reception.

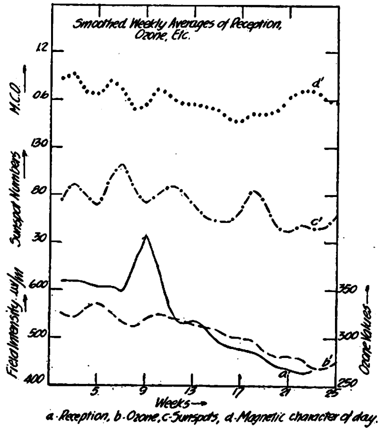

Fig. 10

As pointed out in the letter in *Nature*, already referred to, the close correlation between ozone and reception on 4 km is striking in view of the great distance of about 8000 km, from Bangalore and Madras to the stations in northwestern Europe, where the ozone values have been observed. The conclusion seems to be warranted that changes in atmospheric ozone partake of the nature of a world phenomenon. That this is not so has been pointed out by Dobson. The reasons given are (a) that ozone values have an annual variation in higher latitudes, (b) that the corresponding variations in lower latitudes are correspondingly small, and (c) that no evidence has been found of any world wide variations in the amount of ozone.

While great care has to be exercised in coming to any definite conclusion on this point, the similarity between the curves is so marked over a comparatively considerable period of time, that the correlation may not be found to be spurious after an examination of further data.

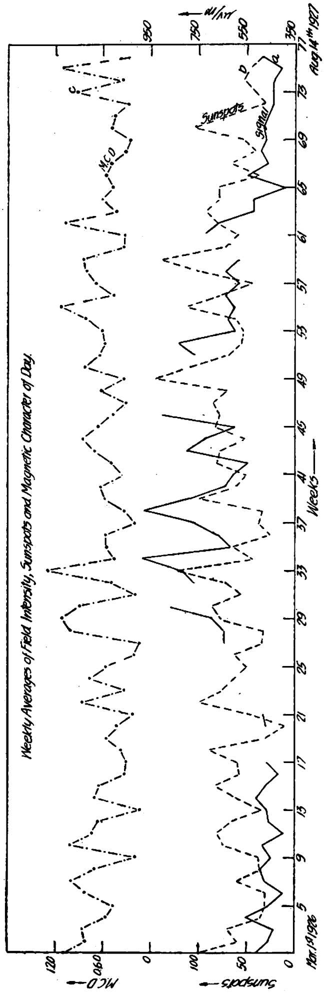
Weekly Averages of Field Intensity, Sunspots and Magnetic Character of Day.
µV/MT
950
750
550
350
77
Aug 14th 1927
73
69
65
61
57
53
49
45
41
37
33
29
25
21
17
13
9
5
Mar. 19th 1926
Weeks
M.C.D
Sunspots
Signal
a
b
c
120
0.60
0
100
50
0
M.C.D.
Sunspots

Fig. 11

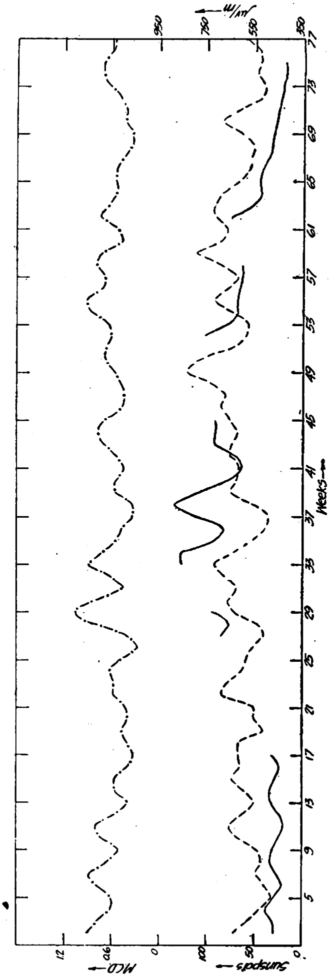
µV/MT
950
750
550
350
77
73
69
65
61
57
53
49
45
41
37
33
29
25
21
17
13
9
5
Weeks
1.2
0.6
0
100
50
0
M.C.D.
Sunspots

Fig. 12

Dobson estimates the height of the ozone layer at about 30 to 40 km. Eckersley's estimate[10] of the height up to which long waves travel is between 40 and 50 km. Hollingworth (loc. cit.) has given the limits as 75 and 90 km. If Eckersley's figure is taken, the existence of the ozone layer at about that height is of very great interest.

What part ozone plays in the ionization changes in the upper atmosphere at periods of sunrise and sunset would be interesting to know; at the equator, where the twilight period is very small, the changes in ionization take place rapidly while in the northern latitudes, where the period is of a few hours duration, there is much more gradualness in the change. Hollingworth[11] has shown that long waves undergo changes in intensity and polarization in a regular cyclical manner at these periods, when the transmitting station is at a moderate distance.

It is to be emphasized that the relation found between ozone and reception requires for confirmation prolonged observations at Bangalore on both ozone and signal intensity before it can be accepted as established.

6. TERRESTRIAL MAGNETISM, SUNSPOTS AND RECEPTION

In studying the variations in the field intensity of Madras in relation to terrestrial magnetism, considerable use has been made of the international numbers for the magnetic character of day. They represent the sum of the character of day figures of the forty-three magnetic observatories all over the world, cooperating under the international arrangement. Variations in declination and horizontal intensity were found more difficult for computation and were not therefore used; and for the study with sunspots, Dr. Wolfer's sunspot numbers have been used. It is to be remembered that the M.C.D. has no real physical basis by itself and from that point of view is perhaps unsuitable for comparisons of this kind.

In Fig. 11, *a*, *b*, and *c* represent from March 1st, 1926 to August 14th, 1927, the respective weekly averages of signal strength, Wolfer's sunspot numbers and the international numbers for the magnetic character of day (M.C.D.) The curves *a*, *b*, and *c* of Fig. 12 give the smoothed values obtained as described earlier. There appears to be no relation of any kind between reception and M.C.D., either of a direct or of an inverse nature. Analysis of daily values and monthly averages too have shown no connection whatever. A day of magnetic storm represented by a high character number is generally unaccompanied by any abnormality of reception even with some reasonable difference in time. The curves of smoothed values too (Fig. 12) show no relation between reception and M.C.D.

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[10] Round, Eckersley, Tremellen, and Lunnon, "Report on Measurements Made on Signal Strength at Great Distances," *Jour.* I. E. E., 63, 933, 1925.

[11] J. Hollingworth, "The Polarisation of Radio Waves," *Proc.* Roy. Soc., 119A, 444, 1928.

Further examination of this point was made by analyzing the readings on the basis of the five selected *D* days of the month. Following the method of Chree and Stagg (loc. cit.), each of these five *D* days of the month is given the number *n*; the succeeding days are $n+1$, $n+2$, etc.; while the days preceding it are $n-1$, $n-2$, etc. Under each number the values of the M.C.D. and signal strength for the corresponding dates are written side by side. The process is repeated for each of the eighteen months, thus giving ninety *n* days and an equal number of $n-1$, $n-2$ etc., and $n+1$, $n+2$ etc., days. The average of the two quantities for each of the days on either side of, and including day *n*, is taken. Thus we get their average daily variation on both sides of the *D* days. Table 1 for April 1927 is given to make the process clear. The *D* days for this month are 9th, 11th, 12th, 14th, and 24th. The readings for the rest of the months go one after another in the respective columns.

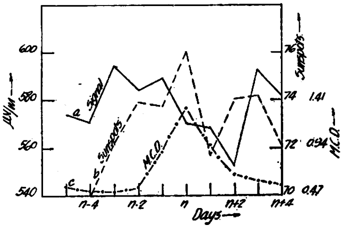

Fig. 13

The averages for the five days before and four days after day *n*—ten days in all—are given as curves in Fig. 13. No evidence of similarity or of any inverse relation is present. While the M.C.D. varies between the wide limits of 0.5 to 1.5, i.e., from a comparatively mild to a highly disturbed day, the maximum variation in sunspots is from 70 to 76, about 8 per cent, and that in field intensity from 550 to 590 μv per m, about 7 per cent. Earlier in the paper, it was seen that the ozone value too exhibited little relation with terrestrial magnetism as represented by the M.C.D. Reference to this will be made again a little later.

$n-4$			$n-3$			$n-2$			$n-1$		
Date	M.C.D.	F.S.	Date	M.C.D.	F.S.	Date	M.C.D.	F.S.	Date	M.C.D.	F.S.
Apr. 5	29	638	Apr. 6	10	640	Apr. 7	37	635	Apr. 8	37	691
7	37	635	8	37	690	9	57	600	10	31	638
8	37	690	9	57	600	10	31	638	11	58	590
10	31	638	11	58	590	12	40	560	13	37	495
20	4	—	21	2	—	22	1	—	23	35	—

n			$n+1$			$n+2$			$n+3$		
Date	M.C.D.	F.S.	Date	M.C.D.	F.S.	Date	M.C.D.	F.S.	Date	M.C.D.	F.S.
Apr. 9	57	600	Apr. 10	31	638	Apr. 11	58	590	Apr. 12	40	560
11	58	590	12	40	560	13	37	495	14	76	578
12	40	560	13	37	495	14	76	573	15	34	580
14	76	573	15	34	580	16	5	580	17	1	790
24	49	—	25	34	—	26	9	675	27	4	—

Reception and Sunspots. In their main variation as shown by curves *a* and *b* of Fig. 11, there appears to be a direct relationship between reception and sunspots, an increase or decrease of sunspots being usually accompanied by a corresponding change in reception. It is to be strongly emphasized that there is little similarity between the curves in the matter of details, due perhaps to other influences on reception. There is also a shift of phase between the two sets of curves of no constant character; and even in the cases of similarity, the proportional changes of amplitude in the two quantities are not constant.

From the curves of smoothed averages (Fig. 12), the general characteristics are seen to be that from March to July 1926, the values are fairly constant varying only a little from the average. From July onward, they are on the increase till March 1927, when there is a steeper increase back to about the same value for the corresponding month of 1926.

7. THE 27-DAY PERIOD

In their analysis (loc. cit.) Chree and Stagg have found a 27-day interval in magnetic disturbances of any type and concluded that there is no definite departure from this period, whatever be the sunspot conditions prevailing. Further, the period of solar rotation in the region of the greatest number of spots is approximately 27 days. This coincidence is of help in the present study of the relation between solar phenomena and radio reception.

The whole period of 18 months has been divided up into twenty groups of 27 days each and the observational data arranged accordingly. The field intensities of the first days in each group are added and the average calculated. The readings for the second days in each group are similarly added and the average taken. The process is repeated for each of the 27 days. Thus a series of average daily values for the period is obtained, giving the mean day to day variation for the 27 days, removing almost all the undesired variations. Similarly we get the figures for sunspot numbers and the M.C.D. These give us Fig. 14 with the curves *a*, *b*, and *c* while *a′*, *b′*, and *c′* give the smoothed values. The coefficient of correlation between sunspots and reception works out to 0.31 ± 0.117, from curves *a* and *b*, Fig. 14.

The agreement between the curves of sunspot and reception even in the matter of peaks and troughs is pronounced. On four of the days there is a phase shift of a day, with sunspots as the earlier occurrence. On the nineteenth day, however, the reverse is true. The ratio of field strength to sunspots for the averages over the whole interval is 8.86. The maximum ratio is 10.6 and the minimum 7.2, giving a maximum variation range of about 20 per cent. In view of the long period considered and the nature of the phenomena, the above is certainly satisfactory.

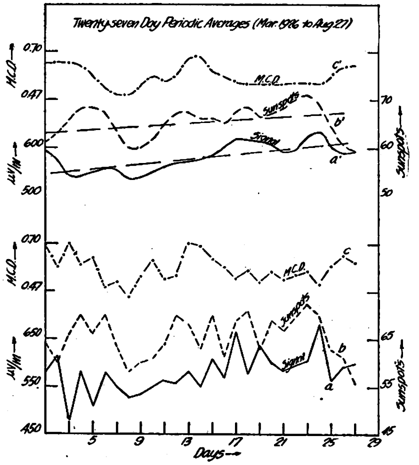

Fig. 14

In addition to the daily variations, a main steady increase in both the quantities is evident at a glance as Fig. 14 will show. This is better brought out in the smoothed curves *a′* and *b′* of Fig. 14. In this connection, it is necessary to point out that while the main general increase is a common feature and there is agreement in the details of peaks and troughs, the proportionate changes of amplitude are not the same right through.

Examining the curves of the M.C.D., there are a few days when the peaks and troughs correspond with those in the sunspot curve, otherwise there is no other feature of any interest.

The 6-Day Period. Curve *a′* (Fig. 14) exhibits what appears to be a six-day period. The peaks occur on the days 1, 6, 12, 17, and 24; and the troughs on the days 4, 9, 14, 21, and 27. No such shorter period is seen in the sunspot curve. This feature is more prominent when the 27-day periodic averages are made out for the 12 months from March 1926 to March 1927, as given in Fig. 15; the coefficient of correlation in this case is much higher, 0.73 ± 0.06.

While the appearance of this period seems definite from the curves, it is interesting to note that Austin has discovered a 9-day periodic variation of intensity with some stations. With the reasonable assumption of the absence of periods smaller than 27 days in solar and terrestrial magnetic disturbances, it follows that the shorter periodic variations of reception, if their existence is world wide, should be due to peculiarities of the locality itself. The Bangalore and Washington observations lead to this inference. What this local influence can be, it is difficult to see, unless it be temperature.

Variations in Amplitudes. With qualitative similarity in the changes in sunspots and reception, there is no constancy in the ratios of the amplitudes of such variations. A small increase in sunspots may result in an increase in reception. But a much bigger change in sunspots may not, and usually is not, accompanied by a proportionate increase in reception. This along with the phase shift of about two days suggests the following explanation.

On the Kennelly-Heaviside layer theory, the main cause of the changes in long-wave signal strength consists in the changes in the conductivity of the upper layers; the variations in ionization at these heights, apart from any radiations due to the earth's surface, are due to the ionizing agents emitted by the sun. These consist mainly of charged or neutral high speed corpuscles and the portion of the electromagnetic radiations which have ionizing properties, such as the ultra violet portion of the spectrum. Of the total emission, the earth gets its tiny share. The only pertinent point about the spots on the surface of the sun is that when some of these come in the direction of the earth, there is a change in the quantity of the ionizing agents reaching the earth. So, if every sunspot was equally effective in regard to the changes in atmospheric conductivity, then Wolfer's numbers would truly correspond with reception in direct proportion. In the absence of it, either all the spots are not equally active or the emission of the high speed particles or of changes in the electromagnetic radiations, due to some of the spots, are directed away from reaching the earth's surface. In such a case, Wolfer's numbers are accompanied by only a little or no change at all in field intensity.

In Figs. 14 and 15, there appears on a few days a lag of about two days between sunspots and signal strength, while for the most part of the 27 days there is full agreement. This can be accounted for on the assumption that the sunspots change only the number of the high velocity corpuscles shot out into the earth's atmosphere.

When the spots are few, if solar activity is violent the velocity of the particles shot out is very high, possibly as much as 1/10 of that of light. Neglecting any retarding influences on the way, these would enter the earth's atmosphere within a few hours and cause changes in the conductivity of the atmosphere, resulting in an almost immediate response in radio field strength variations. But when the velocity of the particles is comparatively low, a much longer time would be necessary to reach the earth's atmosphere. Even if their numbers were high, the velocity being low, the effective change in ionization of the earth's atmosphere in the region where long waves are supposed to travel is not great. The lag of two days would mean an average velocity for the particles of about 1/360 of the velocity of light. Further, the diffusion process from the heights at which these high velocity particles enter into the earth's atmosphere to the heights above the surface of the earth where long waves are believed to travel, is likely to take considerable time.

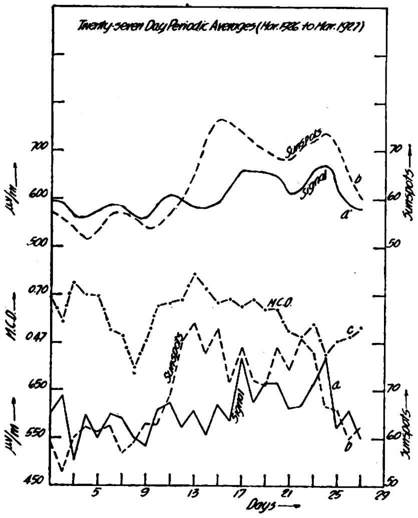

Fig. 15

8. CERTAIN DIFFERENCES BETWEEN LONG- AND SHORT-WAVE TRANSMISSION

It has often been pointed out that there are many differences in the transmission of long and short waves. Of these the present paper attempts an explanation of one or two only, mainly concerned with the subject matter of the paper.

We have, first of all, the phenomenon that during a time when short-wave propagation shows great abnormality, lower frequencies may exhibit none. A striking instance of this was in October 1926, when short-wave communication the world over broke down while long-wave reception was hardly affected. A milder instance was on July 8, 1928, when for over 24 hours, short-wave communication was very difficult.

Secondly, there is the occasional lag of about two days, at the most, between long-wave reception and sunspots, whereas short-wave propagation indicates almost instantaneous response.

Thirdly, we have the possibility of a true physical relation between long-wave reception and atmospheric ozone. With short waves, however, no such relation has been noticed. These three are not independent of each other and any explanation should consider them together.

Now, apart from the details of the constitution of the upper atmosphere about which complete definite knowledge is lacking, there are two points which can be safely taken as correct.

(a) At comparatively great heights, the density is so low that free electrons can exist with a long mean path; at lower heights, say 50 km, density is higher, and free electrons cannot exist, so that there are only ions of very much less mobility.

(b) All the measurements agree that short waves penetrate into greater heights than long waves. For the former, varying figures have been given, from 100 to 250 or 300 km. For long waves the highest figure given is between 75 and 90 km.

The suggestion is that short waves are propagated in an absorbing refractive medium consisting of free electrons, while for the transmission of lower frequencies the conductivity of the lower layers of less mobile ions is responsible.

The fundamental equation for the passage of a radio ray of velocity $\omega = 2\pi f$ in a cloud of electrons of density N and collisional frequency n is

$$I_z = NeV_z = \frac{Ne^2E(kn - i\omega)}{m(kn^2 + \omega^2)}$$

where e and m represent the charge and mass of the electron, while E the electric vector of the plane polarized ray is parallel to z. The displacement or condenser portion of the current is represented by the second term, while the first one represents the collisional momentum of the electrons, the two having a 90-deg. phase difference. The coefficient of collision differs little from unity. While the second term helps in the propagation of the wave, the first one robs the energy in the wave and thus attenuates it giving the familiar exponential term $\epsilon^{-\alpha d/\phi(u)}$ in its most general form. When $\omega \gg n$, the attenuation is small and the electrons execute a number of oscillations before colliding, a case of refraction with negligible absorption. But when the frequency of collision is high, that is, when $n \gg \omega$, the ray is attenuated rapidly.

Short-wave observations seem to agree with the above rough picture of propagation; for the increase in the density of the electron cloud due to increased sunspots or to any other agency is always attended with poorer short-wave reception, as demanded by theory. Larmor[12] suggested that the magnetic storm of October, 1926, might be attributed to the incursion of electrons—not necessarily a large number—into our atmosphere, tending to twist out all the usual ray paths. This seems hardly possible in view of the world wide blocking of short-wave communication. A more satisfactory explanation is that of absorption due to the additional electrons, as represented by the first term of the above equation. The number of them, while sufficient to have a pronounced effect on short-wave reception may not in any way affect long-wave transmission even if the velocities of the electrons were high; the smallness in their numbers would effectively prevent the ionization due to them from penetrating into the lower regions of long waves.

As remarked earlier, the effect of the spots on the sun coming in the direction of the earth is to cause a change in the ionizing agents of the earth's atmosphere, that is, in the high speed particles and in the electromagnetic radiations. The latter make their presence felt by the corresponding changes of ionization occurring almost immediately by variations in short-wave reception. The effect on long-wave transmission depends upon the magnitude of the change in the radiation. If of sufficient intensity to penetrate into the lower heights, there will be variations on long-wave propagation too.

In the case of high speed corpuscles, the number and the velocity of the particles determine the change. A large number with small velocities, besides taking a long time to travel to the earth's atmosphere, may not cause any serious changes in reception.

These considerations and the correlation between the Bangalore measurements on 4 km, with the ozone content of the atmosphere, give an indication that electrons are responsible for short-wave propagation while the less mobile and heavier ions take part in long-wave transmission. As though to support this view, we have the tantalizing estimate of the ozone layer at about 30 or 40 km. High sunspot numbers, poor short-wave reception, improved long-wave transmission seem to go together with increase in the ozone value of the atmosphere and reduced temperature. The suggestion made above is qualitative. A quantitative treatment with experimental data for an extended period will no doubt take into consideration the action of the earth's field. Considerable data and analysis are necessary before some of the results of the paper can be accepted as established and beyond doubt.

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[12] J. J. Larmor, "Magnetic Storms and Wireless Communication," *Nature*, **118,** 662, 1926.

I am indebted to Sjt. C. Seshachar of the Bangalore Observatory for the temperature and pressure readings at Bangalore, and to Sjt. A. A. Narayana Iyer for the temperature readings at Madras. For the magnetic data, I have to thank the late Dr. C. Chree, the Astronomer-Royal and D. K. R. Ramanathan. To Dr. G. M. B. Dobson I owe the ozone values on which that portion of the paper is based. I am grateful to Dr. M. O. Forster for kind permission to publish the paper in the PROCEEDINGS. It is with pleasure that I acknowledge the interest that Prof. J. K. Catterson-Smith has always taken in these measurements which were made with his encouragement at Bangalore.

Addendum

Information so far obtained since January, 1929, from measurements of ozone at lower latitudes shows that its seasonal variations are quite small. Also the only figure, that due to Eckersley, for the height of the Heaviside layer, which at all approximated to that of the ozone layer, has been revised to lie between 80 and 100 km. This, of course, is the usually accepted view for long waves.

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THE SIGNIFICANCE OF OBSERVATIONS OF THE PHASE OF RADIO ECHOES*

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Summary—*A method of observing the phase of radio echoes has been developed by Messrs. Tuve and Hafstad. The present note is intended to show to what extent the phase can be expected to be constant throughout the echo if the frequency-dispersion of the Kennelly-Heaviside layer is taken into account. It is shown that for reflections of 4000-kc waves with a small retardation (effective height of 200 km) the phase can be expected to be the same for the whole echo and that in this case the observed phase is a measure of the optical path. For echoes which spend a longer time in the layer (effective height of 1800 km) the phase should not be constant and the average phase is not expected to be a measure of the optical path.*

It is shown that by measurements on reflections with low retardation the ratio between the changes of (1) the equivalent height for interference and (2) the effective height for echo retardation is a measure of how much of the change is due to the layer moving as a whole and how much is due to a redistribution of electron-densities through the layer. A compression or expansion of a layer having an electron-density increasing in proportion to the height above the lower boundary should give a result by the interference-method of approximately one-third the value obtained by the echo-retardations.

It is shown that the broad echoes observed at night by Hafstad and Tuve are in all probability due to multiple echoes and not to frequency-distortion in the sidebands of the emitted pulses.

THE information received by means of radio about the height of the Kennelly-Heaviside layer has heretofore been principally from the following sources: (a) skip-distance measurements according to Taylor and Hulburt;[1] (b) wavelength variation measurement;[2] (c) measurement of the angle of the downcoming wave;[3] (d) observation of the echoes of short pulses.[4] It has been shown by Appleton[5] and Kenrick[6] that (b) and (d) are equivalent in the sense that they both give the time which a group of waves spends in travelling from the transmitter to the receiver. A combination of the two is therefore incapable of giving more information than either by itself, although of course the multiplicity of paths and fading of components is seen more clearly by means of (d). Method (c) is for short waves equivalent to

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* Dewey decimal classification: R113.6. Original manuscript received by the Institute, May 29, 1929.

[1] A. H. Taylor and E. O. Hulburt, *Phys. Rev.*, **27**, 189, 1926.
[2] E. V. Appleton and M. A. F. Barnett, *Proc.* Roy. Soc., A, **113**, 450, 1927.
[3] E. V. Appleton and M. A. F. Barnett, *Proc.* Roy. Soc., A, **109**, 621, 1925.
[4] G. Breit and M. A. Tuve, *Phys. Rev.*, **28**, 554, 1926.
[5] E. V. Appleton, *Proc.* Phys. Soc., **41**, 43, 1928.
[6] G. W. Kenrick and C. K. Jen, Proc. I. R. E., **17**, 711; April, 1929.

(b) and (d) since the tangents to the ray at the earth determine the equivalent path for echo-retardation. For longer waves it may be possible to obtain additional information from (c), particularly for the circularly polarized components. Short waves, however, penetrate the layer better than the long, and it is especially desirable to obtain additional data by means of them. Method (a) serves as a useful check but on account of the complexity of the phenomena involved gives average values only and is also concerned primarily with the angles which the rays make with the vertical.

For short waves these methods give us essentially $\int(1/\mu)ds$ where μ is the refractive index and the $\int$ is taken from end to end of the signal path. The echo-interference method suggested by the writer and carried out by Messrs. Tuve and Hafstad[7] is intended to give information about $\int\mu ds$, the optical path of the ray. The method consists in introducing an e.m.f. in the receiver due to the continuously going crystal oscillator used for frequency control in the transmitter. As the echoes arrive their effect depends on their phase with respect to the phase of the crystal. If, for instance, the echo and the crystal are in phase the effect on the receiver is large; if they are out of phase the combined effect is small. It is shown below that to a good approximation the relative phase depends on $\int\mu ds$. If the echo-interference and the echo-retardation are measured simultaneously a comparison of changes in the equivalent path for interference and the effective path for echo-retardation gives at least an idea of the manner in which the change in the layer takes place, for instance, of whether the layer moved as a whole or whether the lower boundary remained fixed, but in the layer the regions of larger electron densities moved up. By means of this method it is possible to detect smaller changes in height than have been observed otherwise. On account of the complexity of the echo-pattern it is important to resolve the pattern rather than to be analyzing fading curves which are the result of the superposition of a large number of interfering echoes. In order to be sure of the interpretation of the results it is of course necessary also to study the polarization of the echoes. In the calculations made below nothing is said about polarization. This omission is not important since polarization can be taken into account later and would needlessly complicate the present discussion, which is concerned with the frequency-distortion due to the sidebands of the pulses used.

Let us consider a set of rays connecting the transmitter T to the receiver R. For every frequency f there is a definite $\int(v(f))^{-1}ds$ where $v(f)$ is the phase-velocity. The disturbance at R due to a pulse at T is

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[7] L. R. Hafstad and M. A. Tuve, Proc. I. R. E., 17, 1786; October, 1929.

$$E=\int_0^\infty A(f)\exp 2\pi i\left(ft-\int(f/v(f))ds\right)fd \tag{1}$$

while the disturbance sent out by T is

$$F(t)=\int_0^\infty A(f)\exp 2\pi ift\ df=g(t)\exp 2\pi if_0t. \tag{2}$$

We let

$$\int(f/v(f))ds=P(f) \tag{3}$$

and expand by Taylor's theorem

$$P(f)=P(f_0)+(f-f_0)P'(f_0)+(1/2)(f-f_0)^2P(f_0)+\cdots. \tag{4}$$

For the present we confine ourselves to the first two terms. Substituting (4) in (1) we have

$$\begin{aligned}E&=\int_0^\infty A(f)\exp 2\pi i(f(t-P_0')-P_0+P_0'f_0)df\\&=F(t-P_0')\exp 2\pi i(f_0P_0'-P_0)=g(t-P_0')\exp 2\pi i(f_0t-P_0)\end{aligned} \tag{5}$$

As long as the first two terms of the expansion (4) are sufficiently accurate we see that the envelope g of the disturbance E is retarded by the amount P_0' while the phase of the vibrations modulated by g is retarded by $2\pi P_0$. According to (3) the phase depends simply on the optical path since

$$P(f)=\int(f/c)\mu(f)ds \tag{6}$$

where $\mu(f)$ is the refractive index at frequency f. The retardation of the envelope, or in other words the arrival of the group, depends on P_0'. By Fermat's theorem the differentiation can be performed inside the integral (4) so that

$$P'=\int(v_g(f))^{-1}ds\ ;\qquad (v_g(f))^{-1}=\frac{d}{df}(f/v(f) \tag{7}$$

where v_g is the group velocity. We conclude that as long as terms beyond the second are negligible in (4) the time of arrival of the group of waves represented by (2) is given by the group-velocity while the phase of the vibrations when they arrive is given by the phase velocity. It is not clear without calculation to what extent the terms disregarded in (4) are negligible. The third term when inserted in the integral of (5) gives on expanding the exponential a corrected value of E

$$E = \sum_{n=0}^{\infty}(-\pi i P_0'')^n(1/n!)$$

$$\int_0^{\infty}(f-f_0)^{2n}A(f)\ \exp\ 2\pi i f t' df\ \exp\ 2\pi i(f_0 P_0' - P_0) \tag{8}$$

where

$$t' = t - P_0'.$$

In order to evaluate the integrals involving $(f-f_0)^{2n}$ we differentiate (2) with respect to t twice and obtain

$$\int_0^{\infty} f A(f)\ \exp\ 2\pi i f t\ df = ((g'/2\pi i) + f_0 g)\ \exp\ 2\pi i f_0 t\ ;$$

$$\int_0^{\infty} f^2 A(f)\ \exp\ 2\pi i f t\ df = (-(g''/4\pi^2) + (f_0 g'/\pi i) + f_0^2 g)\ \exp\ 2\pi i f_0 t$$

whence combining with (2)

$$\int_0^{\infty}(f-f_0)^2 A(f)\ \exp\ 2\pi i f t\ df = -(g''/4\pi^2)\ \exp\ 2\pi i f_0 t$$

and repeating the process n times

$$\int_0^{\infty}(f-f_0)^{2n}A(f)\ \exp\ 2\pi i f t\ df = (-1/4\pi^2)^n \frac{d^{2n}g}{dt^{2n}}\ \exp\ 2\pi i f_0 t$$

substituting this in (8)

$$E = \sum_{n=0}^{\infty}(iP_0''/4\pi)^n(1/n!)\left(\frac{d^{2n}g}{dt^{2n}\,t'}\right)\ \exp\ 2\pi i(f_0 t - P_0). \tag{9}$$

The factor in front of the exponential is complex, and therefore the phase of the vibrations is affected. The effect is in general different throughout the received hump. In the special case of derivatives of g higher than the second vanishing we have a simplified form

$$E = (g + (iP_0''g''/4\pi))\ \exp\ 2\pi i(f_0 t - P_0). \tag{9'}$$

The phase correction to be applied to (5) is

$$\theta = \tan^{-1}\ (P_0''g''/4\pi g) \cong P_0''g''/4\pi g \tag{10}$$

measured in radians. A special case of an envelope g is a sine-curve of period $2T$, the "on time" being T. For this case $(g''/g) = -\pi^2/T^2$ and the phase correction is

$$\theta = -\pi P_0''/4T^2. \tag{10'}$$

Let us examine this formula for the case of a uniformly increasing density of electrons and a somewhat simplified dependence of the refractive index on the frequency: $\mu = 1 - 2by/f^2$ where y is the height above the lower boundary of the layer. For normal incidence we find $P(f) = 2f^3/3bc = 4hf/3c$, where $h = f^2/2b$ gives the height to which the signal rises above lower boundary. Letting h' be the effective height to which the signal rises above lower boundary as measured by the echo-method, we have $P' = 2f^2/bc = 2h'/c$ so that $h' = f^2/b$. The effective height h' is in this case just twice the actual height h. The equivalent height for interference is a still third and different quantity $h_1 = 2h/3 = h'/3$. The error in the phase due to variation in group velocity is by (10′) $\theta = -\pi h'/cfT^2$. If $h' = 100$ km, $T = 10^{-4}$ sec., $f = 4 \times 10^6$ cycles, corresponding to the conditions of the experiment we have $\theta = -1.5$ deg. This is a small and at present hardly measurable error. In this estimate f has been underestimated on purpose and T is probably also underestimated. Taking h' to be 100 km amounts to supposing that the first reflection of approximate effective height 200 km is due to a 100-km path (one way) in the lower un-ionized atmosphere and an equal path in the Kennelly-Heaviside layer. For the first reflection h' is surely less than 200 km so that the maximum error in the phase is for this case 3 deg. For reflections with higher retardations, say for effective heights of 1800 km, we expect proportionately higher errors which in this case have to be obtained by means of (9).

Since we have disregarded terms higher than the third in (4) we have still an error in (9), but for practical purposes this error is small. The fourth term introduces in (5) a correction factor $\exp(\pi i/3)(f-f_0)^3 P_0'''$. For the special case just considered this means $\exp(4\pi i/3c)h'(f-f_0)^3 f^{-2}$. If $f - f_0 = 10^4$ the angle is 0.005 deg. At the most we are concerned with $f - f_0 = 3 \times 10^4$ in which case the angle is 0.13 deg. The effect of this term on the resultant vibrations is presumably much less than that, because the first odd term in (4) involving $(f-f_0)^{-1}$ has no effect on the phase. We can draw another conclusion from (9). It shows that if g and its derivatives vanish there is no signal. This means that to within the first three terms of (4) the arrival of the humps is accurately given by the group-velocity and that their width is the original width of the pulse sent out by the transmitter. It is therefore very unlikely that the broadened reflections observed at night by Hafstad and Tuve[8] are due to distortion of the signals by the dispersing action of the ionized atmosphere. They are rather to be explained as the result of the superposition of several echoes. This conclusion is especially supported by an experiment performed during one of the night tests. The carrier-frequency was changed by 20 kc. No observable change in the retardation of the broad echoes was found. If the variations of group-velocity in the side-bands were responsible for the width, this change of all the frequencies concerned would have produced a measurable change in the effective height.

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[8] L. R. Hafstad and M. A. Tuve, Proc. I. R. E., **17**, 1513; September, 1929.

The law $\mu^2=1-2by/f^2$ is of course only approximate. The conclusions we have reached by means of it are essentially correct. Instead of taking the electron-density to be proportional to the height y above the lower boundary of the layer, we shall suppose that it is proportional to y^n where n is a positive constant, and we also attempt to take into account the magnetic properties of the medium by letting

$$\mu^2=1-a^n y^n/(f^2-ff_1) \tag{11}$$

where f_1 may be the magnetic absorption frequency or its negative. This gives

$$P(f)=2f(f^2-ff_1)^{1/n}q_n/ac=2fhq_n/c \tag{12}$$

where

$$q_n=\int_0^1(1-x^n)^{1/2}dx,\quad h=(f^2-ff_1)^{1/n}/a$$

As before h is the height reached by the signal. The equivalent height for interference is

$$h_1=hq_n. \tag{13}$$

The effective height for echo measurements is

$$h'=h_1[(1+2/n)f-(1+1/n)f_1](f-f_1)^{-1} \tag{14}$$

while the phase-error is

$$\theta=-(\pi h'/2T^2c)\{(1/nf)+(1-n)/(n(f-f_1)) \\ +(1+2/n)[(1+2/n)f-(1+1/n)f_1]^{-1}\}$$

Relation (14) shows how the effective height for echo-retardation h' is connected with the equivalent height for interference h_1. If n is large $h'=h_1$, while if n is small h' is larger than h_1. This of course is to be expected because for large n the condition of specular reflection is approached, while for small n the rays are gradually refracted. The phase error may become unusually large if $f=f_1$ or if $(n+2)f=(n+1)f_1$. In the region of 4000 kc neither of these conditions takes place and the estimate made with the simplifying assumptions is essentially correct.

As a working hypothesis it is probably satisfactory to take $n=1$ in the layer and to suppose that the lower boundary is at some height h_0

above the ground. If changes in height as observed by the two methods are equal the interpretation is that the layer moves as a whole. If the interference method gives about one-third of the echo retardation all of the change is likely to have taken place as a redistribution of electrons in the layer, resulting in a compression or expansion of the layer as a whole with a fixed lower boundary.

On account of the sensitivity of phase to the magnetic field of the earth it may be possible to detect a correlation between the observed phase-changes and changes in the field.

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INDUCTANCE AS AFFECTED BY THE INITIAL MAGNETIC STATE, AIR-GAP, AND SUPERPOSED CURRENTS*

By
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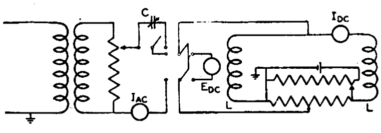
The object of this paper is to present information regarding the manner in which the inductance of iron-cored coils depends upon the initial magnetic state, the length of the air-gap, and the amount of direct and alternating current flowing in the winding. Important contributions have been made by Ball, Chubb, Spooner, Prince, Hanna, Morecroft, Howe, and others, but considering the importance of the subject it is surprising that there are so few data or curves in the literature showing the performance of inductance coils under widely varying conditions. It is hoped that the results here reported will be of interest and value to those who use or occasionally design choke coils or audio-frequency transformers, although no attempt was made to simulate closely the constants of either.

Method of Measurement

The measurements, with the exception of a few with the ballistic galvanometer, were made with the constant impedance method[1] on account of its simplicity, precision, ease of operation, and the fact that it is particularly adapted to the measurement of inductances from five henries up to large values and requires only a small standard condenser which is usually at hand. Since the details of the method are fully covered in the paper referred to, only a brief explanation will be given here.



CONSTANT IMPEDANCE MEASURING CIRCUIT

Fig. 1

Two similar coils are connected in parallel to provide a local path for the direct-current magnetizing component so as to keep it out of the

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[1] H. M. Turner, "The Constant Impedance Method for Measuring Inductance of Choke Coils," Proc. I.R.E., **16**, 1559; November, 1928.

To settle the question, Mammoth Cave, Kentucky, was selected for experiments, and it proved a fortunate choice. This cave has not been electrified for illumination, and the managers were good enough to remove telephone wires. There are no conductors whatever in the cave except here and there short lengths of iron handrails. The exits are filled by and sealed with the waters of Echo River and the River Styx, which flow into Green River. A few years ago a new entrance was found and some electric wiring introduced; this is four miles away overland and eight miles away by cave. A careful survey was made both under and above ground so that corresponding points could be located and levels known. Most of the work was carried out above River Hall, with 300 ft. of overburden,—about 200 ft. of Mammoth Cave limestone capped by 100 ft. of Cypress sandstone. The resistivities of these rocks *in situ* were ascertained by electrical prospecting to be of the order of 10,000 to 20,000 ohms per cm^3.

BROADCASTING WAVELENGTH

The words and music from Louisville, Nashville, and Cincinnati were received on a No. 26 portable Radiola superheterodyne six-tube set, and their bearings were taken with the small loop which is part of the apparatus. These signals were traced into the cave from the entrance and were lost at 500 ft. from the mouth, where the overburden is 150 ft. Yet under the Mammoth Dome, 1,000 ft. in the cave, capped by an ascertained thickness of 75 ft., mostly sandstone, the signals were clear enough, and these were traced along a rather narrow tunnel for 300 ft., when the thickness was 150 ft.

On a later occasion Messrs. Joyce, Barlow, and Kidd took a 300-ft. aerial and gave three turns of it round the loop, so as to form a close coupling, and led the other end to the ground. In this manner there was at night good audible broadcasting in River Hall 300 ft. below the ground, and more than 1,000 ft. from the entrance at a point approached by the famous tortuous passage well named the Corkscrew. On the next night, signals and speech were also audible at Echo River, which is reputed to be 350 ft. underground. The conclusion is irresistible: waves from distant stations can be detected under 300 ft. of sandstone and limestone—waves which do not come through the entrance, and which do not pass along conductors.

LONG-WAVE STATIONS

Morse signals from about six stations with bearings approximately N. 65 W. (Long Island and others) were heard on a loop with a model RE low-frequency receiving equipment, consisting of an antenna coupling unit, a radio-frequency amplifier, a low-frequency receiver, and a tuned audio amplifier. These signals were clear enough to hear, but their intensities were hard to measure, partly owing to static. The wave fronts appeared to be mainly vertical, and the waves must therefore have been travelling through rock rather than down from above. The reception was again at River Hall, 300 ft. below the surface plateau near the camping ground of the Mammoth Cave Hotel. The bearings of these signals obtained above and below ground were nearly identical. This picture of a wave, with its front mainly vertical, travelling through the rock (just as through the air), is rather novel. We could pick up these unidentified stations with three turns of wire on a 4-ft. square frame, and we obtained bearings therewith.

HORIZONTAL WAVE FRONTS

A circular horizontal transmitting loop, 100 ft. in diameter, was placed on the ground, and the insulated wire could be tapped so as to give one to ten turns as required. A rectangular receiving coil of three turns (40×10 sq. ft.) was placed on the floor of the cave at River Hall (300 ft. down). Currents of known magnitude of about half an ampere were introduced into a single turn of the loop, using a gasoline engine generator to excite an oscillator. The high-frequency loop current was modulated by applying a 500-cycle a-c potential on the oscillator plates. The frequency was varied by steps of 10, from 20 to 110 kc. The received voltage on the coil was measured in microvolts, and averaged about four, suggesting a flux of 10^{-9} lines per sq. cm in the cave, due to half an ampere in the loop above ground. It was a most remarkable fact that the 20- and 30-kc frequencies boomed out loudly as compared with the 40- and 100-kc frequencies. Measurements made above ground on the same frequencies showed no such emphasis, and there is a strong suggestion that these frequencies (20 to 30 kc) pass with markedly less absorption through the rock. This question must stand over for a fuller investigation. The selectivity did not appear to be a function of the receiving instrument.

AUDIO FREQUENCIES

When 500-cycle frequency alternating current was impressed on the full ten turns of the 100-ft. diameter circular loop, powerful signals were received with headphones both with and without a three-stage amplifier, using a 400-turns 3×2 sq. ft. rectangular coil, at depths of 100 and of 300 ft. Of particular interest is the fact that detection was readily made without amplifier in the Mammoth Dome, 900 ft. on an inclined line from the horizontal loop. We conclude that the electromagnetic effects of a 500-cycle frequency passed through 900 ft. of continuous rock. It is intended to publish a full report on these experiments in due course.

measuring circuit. With the double-pole switch thrown to the right, adjust the lower sliding contact, Fig. 1, until there is no battery potential between the lines to the parallel inductances as indicated by a low reading voltmeter. This is particularly important when using large values of direct current, for a relatively small unbalanced voltage from the battery might burn out the line instrument or at least introduce slight errors. Throw the double-pole switch to the left or operating position and proceed as follows: with the single-pole switch to the left measure the alternating current in the parallel inductive circuit by means of a thermocouple instrument, electron-tube voltmeter or otherwise, then throw the switch to the right, connecting the standard condenser in series with the line, and starting with a low value of capacity increase it until the line current is the same as before. The impedance is the same for the two positions of the switch, and should be carefully verified by throwing it first to one and then the other, adjusting the capacity until the permanent current is the same.

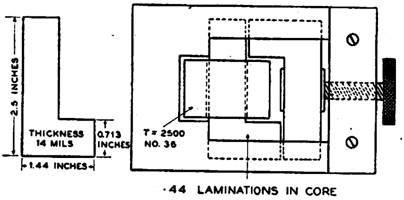


Fig. 2

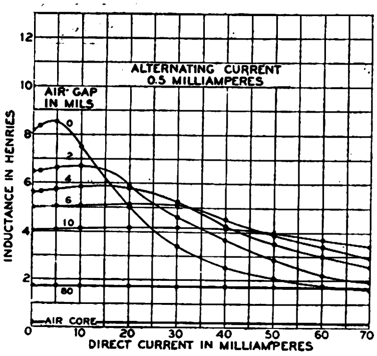


Fig. 3

Since the impedance is the same and no change has been made in the resistance then

$$X_L = X_C - X_L, \qquad \text{or} \qquad X_C = 2X_L \tag{1}$$

that is, the capacity is half that required for resonance at the fundamental frequency and twice that for resonance to the second harmonic, so harmonic components will have negligible effect on accuracy. The adjustment is made on the steep portion of the resonance curve where the critical value of capacity to satisfy equation (1) may be accurately determined. If the line current is too small with the condenser in the circuit the capacity should be increased or vice versa. For critical capacity the permanent line current is the same for the two positions of the single-pole switch, but will lag in one case by θ degrees and lead by θ degrees in the other. The inductance of the two coils in parallel is found from the equation

$$L=\frac{1}{2w^2C} \tag{2}$$

and that for one of the coils by

$$L=\frac{1}{w^2C}\,. \tag{3}$$

The transient currents due to switching will be limited by the inductance and are of no importance, however, if preferred balance may be obtained without opening the circuit. This is done by adjusting the capacity until the voltage across the inductive circuit is the same as that of the line as determined by an electron-tube voltmeter.

MAGNETIC CIRCUIT

The magnetic circuit is shown in Fig. 2 and consists of two L-sections each having 44 laminations 14 mils in thickness, the material being 3.5 per cent silicon steel. The dimensions are: length 16.3 cm, cross-section 2.82 sq. cm and volume 46.2 cu. cm or 6.42 in., 0.437 sq. in. and 2.8 cu. in., respectively. Where the length of the air-gap is

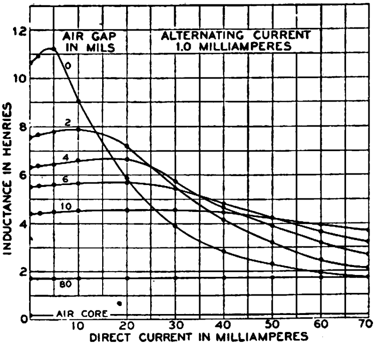


Fig. 4

given it will be the measured value of the two in series. The coil has 2500 turns of No. 36 wire.

CONDITIONS AND MEASUREMENTS

The first series of measurements, the results of which are plotted in Figs. 3 to 13 inclusive, were made under the conditions of practical use, that is, without first demagnetizing the core. As a matter of convenience and to confine the study to the variation of inductance as determined by the magnetic circuit, all measurements in this series were made at 60 cycles. Data were taken over a wide range of conditions: the air-gap being varied from zero to 80 mils and then the core was entirely removed; the direct current was varied from zero to 80 ma and the alternating from 0.5 to 10 ma effective. In order to see more clearly the effect of these various factors families of curves were plotted with each of these quantities as the independent variable. In some respects it would have been better had ampere-turns per unit length of magnetic path or gilberts been used, but with varying air-gap and saturation much of the advantage that would result in the case of complete magnetic circuits would have been lost due to the uncertainty of the portion to be assigned to the air-gap.

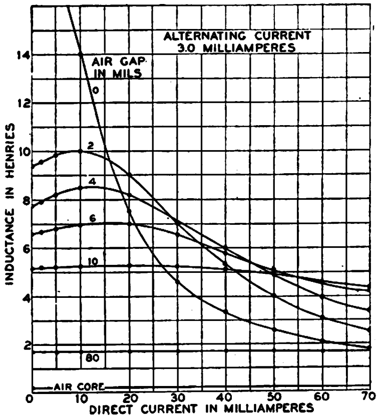


Fig. 5

The second series of measurements, made with the ballistic galvanometer, includes the determination of the inductance for butt and lap joint magnetic circuits where the core was first demagnetized before making measurements and where the core was not demagnetized; normal hysteresis loops for butt and lap joint and the slope of the minor hysteresis loops as determined by the way a given magnetic state is approached, that is, by the past magnetic history.

The third series of measurements was made of the primary inductance of an audio-frequency transformer over a wide range of values of direct and superposed alternating current. These measurements were made with the constant impedance method at 60 cycles to avoid effects that would be present at higher frequencies, such as distributed capacity; however, these effects are small up to a few hundred cycles.

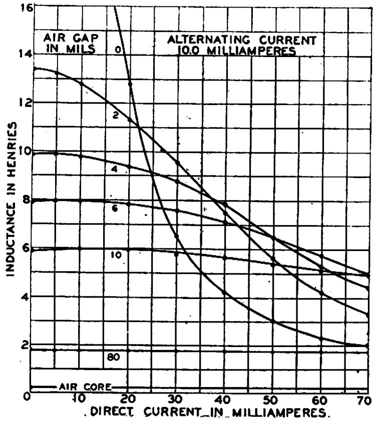


Fig. 6

Discussion of Curves

In analyzing these curves the information regarding the number of turns and core dimensions should be kept in mind. For direct currents up to approximately 12 ma the inductance is greatest for zero air-gap and decreases as the length of the gap increases, as shown in Figs. 3 to 6 inclusive. However, as the direct current increases this order is gradually changed until it is finally completely reversed, except

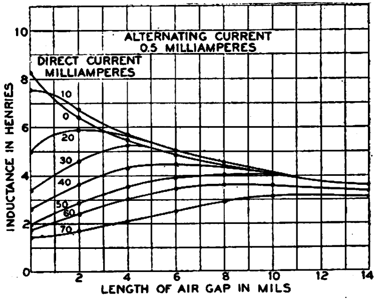


Fig. 7

for the 80-mil gap which falls below the other curves throughout the entire range, being less than half that for the 10-mil gap.

Inductance is usually expressed in terms of the incremental permeability, length of path in iron, length of path in air, and other factors. It is thought that the use of the term incremental reluctance will simplify the discussion and give a clearer understanding of the action taking place. The inductance is given by

$$L = n\frac{d\phi}{di} = \frac{0.4\pi n^2}{R}10^{-8} \text{ henries} \tag{4}$$

$$= \frac{k}{R} = \frac{k}{R_i + R_a} \tag{5}$$

where R_i, the incremental reluctance of the iron, varies with the constant magnetizing component due to saturation of the iron and R_a is that of the air path.

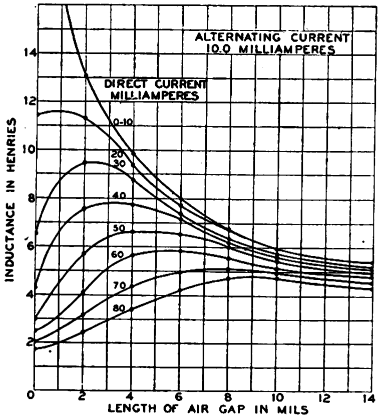


Fig. 8

Overlooking for the moment some secondary effects, the inductance for zero air-gap and low values of direct current is determined largely by R_a, which is appreciable for butt joints even though the two parts of the core are clamped together. R_i is small due to the law saturation of the core.

If the reciprocal of the inductance obtained from Fig. 3 for zero direct current be plotted against the length of the air-gap the curve is practically linear from zero up to 10 mils; however, above this value the reluctance increases less rapidly due to the increase in cross-section of the air path and finally becomes constant. The fact that the curve is linear down to zero indicates that R_a still constitutes the major portion of the reluctance even though there is no apparent air-gap. In order to obtain a measure of the remaining effective air-gap, the core was next built up with a lap joint and the inductance found from which the reluctance was calculated, its value being the ordinate to the dotted horizontal line. The corresponding abscissa is not known; however, it may be approximated by extending the solid line until it intersects the dotted line as shown in Fig. 13, which indicates an eqiuvalent air-gap for the butt joint of approximately 4 mils over that for the lap joint. In the vicinity of the intersection the incremental reluctance of the iron becomes of increasing importance. This was checked by using data obtained from a demagnetized core, and the results were in essential agreement.

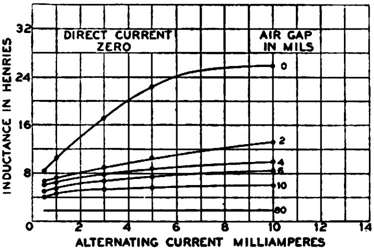


Fig. 9

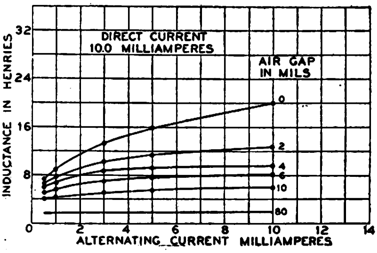


Fig. 10

With a large direct-current component the saturation is such that the incremental reluctance of the iron path exceeds that of the air path, in which case an increase in the length of the gap reduces the constant flux and thereby greatly decreases the incremental reluctance of the iron, that is, R_i decreases faster than R_a increases, and the inductance is larger. This is clearly shown in Figs. 3 to 6, where the inductance for a 10-mil gap and 70 ma direct current is more than twice that for zero gap; however, a point is finally reached where the reverse is true, for example, at the same point the inductance for an 80-mil gap is less than half that for the 10-mil gap. It should be noted that for 0 and 80 mils the inductance is the same, while if the iron core is entirely removed it is reduced to 0.1 this value.

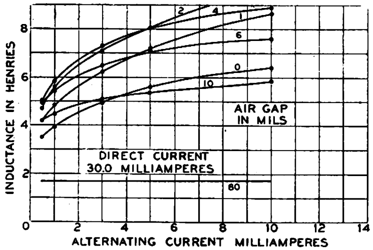


Fig. 11

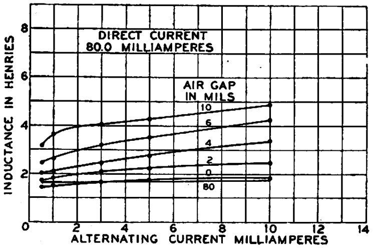


Fig. 12

The inductance increases with the alternating current, indicating the importance of knowing the alternating current as well as the direct current at which the measurements are made if they are to be significant. Where measurements are made without first demagnetizing the core the inductance increases or tends to increase with direct current up to a maximum and then falls off. The greatest increase is for zero gap, and is of the order of a few per cent, say 6 or 8. As the

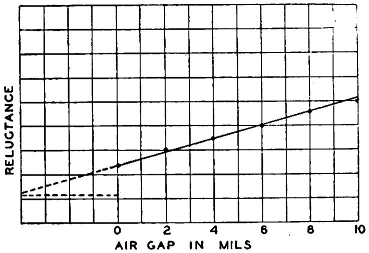


Fig. 13

air-gap is increased the maximum occurs at higher values of direct current, but the percentage becomes less and less.

From a theoretical point of view it might be argued that the core should first be demagnetized, in which case the inductance always decreases with an increase of direct current, but in practice the cores are seldom demagnetized before being used and the inductances may differ widely in some cases. However, for some purposes this procedure is to be recommended.

Figs. 7 and 8 show the effect of air-gap upon the inductance for different values of direct current. Figs. 9 to 12 show the variation as a function of the alternating current. Fig. 13 has already been referred to.

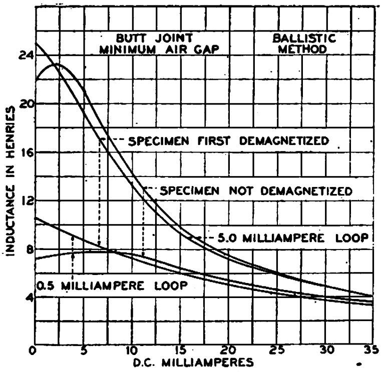


Fig. 14

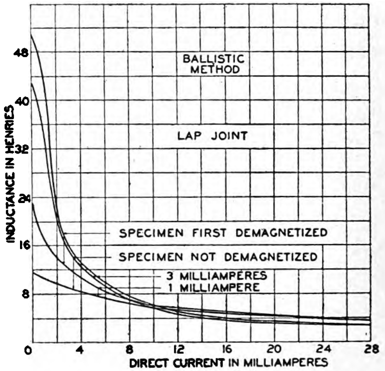


Fig. 15

SECOND SERIES—BALLISTIC METHOD

The difference in inductance obtained from magnetized and demagnetized specimens was investigated, and the results for butt and lap joints are shown in Figs. 14 and 15. For zero direct current the magnetized core gives less inductance, and for small minor loops very much less than for the demagnetized specimen. In some cases this general relation is maintained for increasing direct current, but in others the curves cross. Unfortunately the minor loops in Figs. 14 and 15 are not the same, but it is evident that the lap joint gives a much higher inductance, or lower incremental reluctance than the butt joint, showing definitely that there is a relatively large equivalent air-gap in the best butt joint. This is further borne out by the hysteresis loops in Fig. 16. Attention is called to the low residual for the butt joint, only about 12 per cent, indicating a large equivalent air-gap, while that for the lap joint is about 37 per cent, giving further support to the statement in connection with Fig. 13 that R_a was the major part of the total incremental reluctance when there was no direct current even though the apparent air-gap was zero.

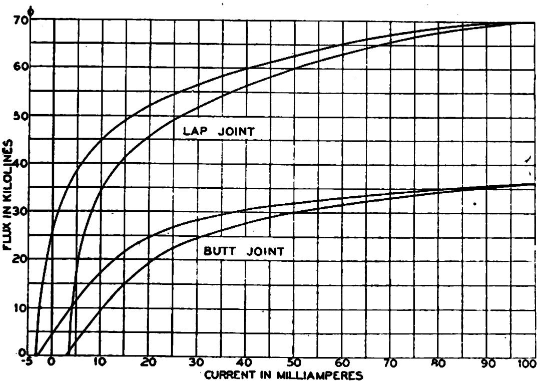


Fig. 16

Had the maximum current been 50 ma in the lower curve, Fig. 16, the upper branch would have joined the one plotted at 18 ma so the maximum value could be considerably reduced below this value and still leave practically the same residual flux.

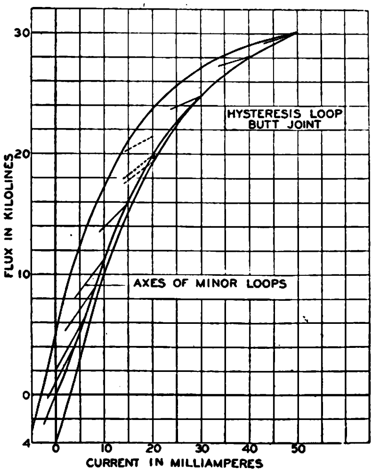
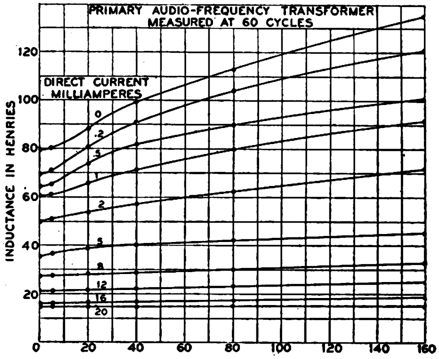


Fig. 17

In Fig. 17 the slope of the minor loops, which is a measure of the incremental permeability, for a demagnetized core decreases with increasing magnetizing current as shown by the solid lines. While they show the correct slope they are not in the proper vertical position and should not have their tips on the initial magnetization curve, but were placed there as a matter of convenience. The dotted lines show the slope and correct vertical position of the minor loops depending on the approach or past magnetic history of the core. The upper one, the slope of which is taken as 100 per cent, was approached from the positive tip of the loop. The middle one was approached from zero on the initial magnetization curve and its slope is 106.8 per cent; the lower one is approached from the negative tip of the normal loop and its slope is 111 per cent.



A. C. MICROAMPERES (EFFECTIVE)

Fig. 18

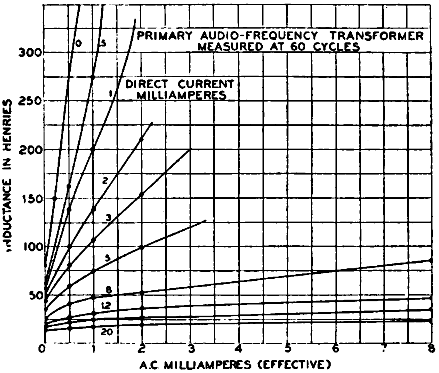


Fig. 19

THIRD SERIES—CONSTANT IMPEDANCE METHOD

The inductance of the primary of an audio-frequency transformer measured over a wide range of direct and alternating currents is given in Figs. 18, 19, and 20. No change was made in the magnetic circuit. The first shows the inductance as a function of the alternating current from extremely low values up to several times the maximum that would ever be encountered. The inductance as a function of the direct current is shown in Fig. 20.

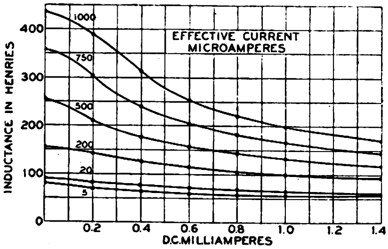


Fig. 20

Thanks are due Lieut. J. W. Green and Mr. H. T. Lyman, Jr., graduate students in Communication Engineering, and Mr. William Sherwood, instructor, for the care exercised in obtaining the data.

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NOTES ON THE DETECTION OF LARGE SIGNALS*

By
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Summary—*The effect of large signals applied to the grid of a detector is discussed. It is shown that signals even as small as fifty mv appreciably affect the tube parameters and influence the frequency distortion. The nature of detector overloading is discussed and overload curves of the plate rectifier are presented.*

MUCH admirable work has been done in developing the theory of rectification in detectors of both types—grid and plate—but almost all of this work has been confined to the case where the radio-frequency signal is quite small—of the order of 50 mv or less. The conditions were assumed to be such that the signal produced no appreciable variation of the tube parameters.

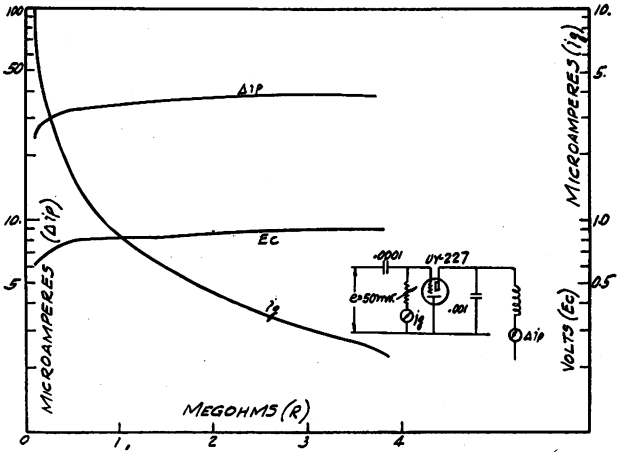

Fig. 1

Actually, it is found, in the case of receivers designed for broadcast reception, where sufficient voltage must be developed to operate a loud speaker at what might be termed "good room volume," that the radio-frequency voltage applied to the detector nearly always exceeds 50 mv, and generally is of the order of 200 or 300 mv, depending of course, on the gain of the audio amplifiers. As the voltage at the detector input is increased, changes in the fidelity of reproduction will be noticed; in addition to this a loss of volume occurs beyond a certain voltage, depending upon the tube parameters.

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In the grid rectifier the audio voltage at the grid is given by[1]

$$e_g = \left(\frac{1}{2} \frac{d^2 i_g}{de_g^2} | Z | \right) m E_0^2 \sin (at + \phi). \tag{1}$$

The quantity in brackets is the detection factor; the term Z is given by

$$Z = \frac{1}{g_g + A_c(d) + A_i(a)}. \tag{2}$$

In these formulas

m = the modulation coefficient;
E_0 = the r-f voltage applied to the grid;
g_g = the grid electron conductance $= 1/R_g$;
A_c = the conductance of the grid-leak, grid-condenser combination.
A_i = the input conductance of the tube due to electrode capacities and plate circuit load.

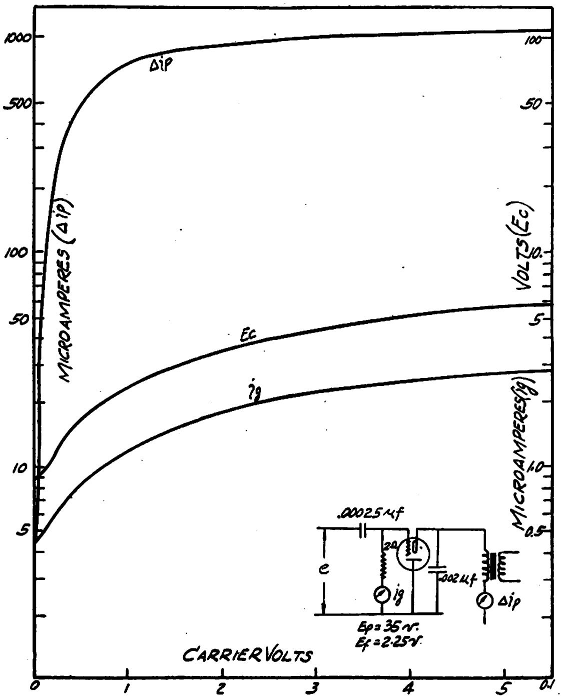

Fig. 2

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[1] S. Ballantine, Proc. I. R. E., **16**, 593; May, 1928.

The values of d^2i_g/de_g^2 are fixed by the constant grid voltage applied through the grid leak. In the case of the UY227 tube, which was used in this work, this initial bias is practically independent of the value of the grid-leak resistance where the latter is returned directly to the cathode. This fact is illustrated in the curves of Fig. 1; a constant signal voltage of 50 mv was applied to the detector; i_g is the grid current; E_c is the product of i_g and the grid-leak resistance, and Δi_p is the change of plate current produced by the signal.

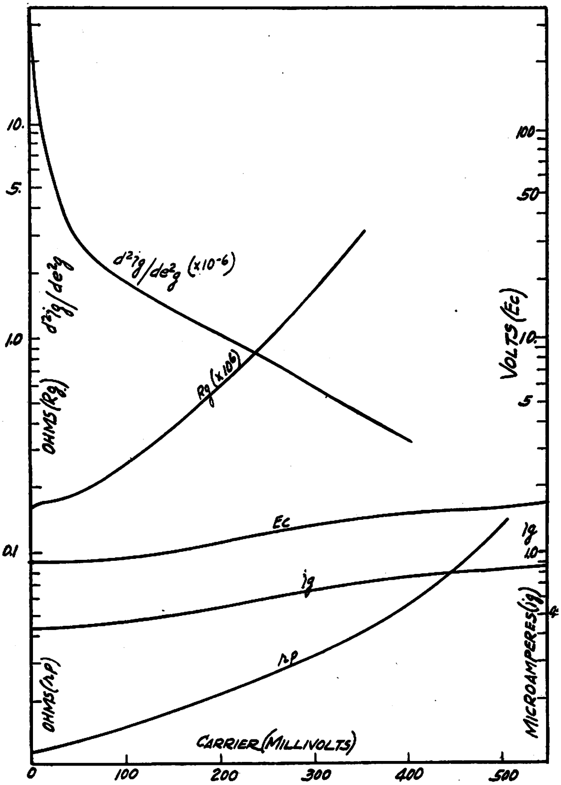

Fig. 3

When, however, the signal voltage was varied, the effects produced were as illustrated in Fig. 2. The grid voltage E_c varies considerably with signal strength, even for small values. Fig. 3 shows the lower parts of these curves magnified, in addition to curves which show the variation of the tube parameters with the signal voltage. These variations are quite large; a 400-mv carrier produces an r_p of 55,000 ohms and an R_g of 2.7 megohms; as compared with initial values of 11,500 and 0.165, respectively. The d^2i_g/de_g^2 changes from 31.5×10^6 to 0.33×10^6.

The effects of these variations are therefore felt at both the input and output of the detector. The effect at the input can be found by applying the values given by the curves of Fig. 3 to formulas (1) and (2), and is indicated by the curves of $|Z|$ and of the detection factor D_g in Fig. 4, where $D_g=\frac{1}{2}(d^2i_g/de_g^2)\,|Z|$.

The ratio of rectified voltage at a modulation frequency of 5000 cycles to the rectified voltage at very low frequencies was calculated and plotted in Fig. 5A. On referring this curve to the R_g curve of Fig. 3, the curve of Fig. 5B was obtained, which indicates the effect of signal strength on the attenuation of the higher audio frequencies.

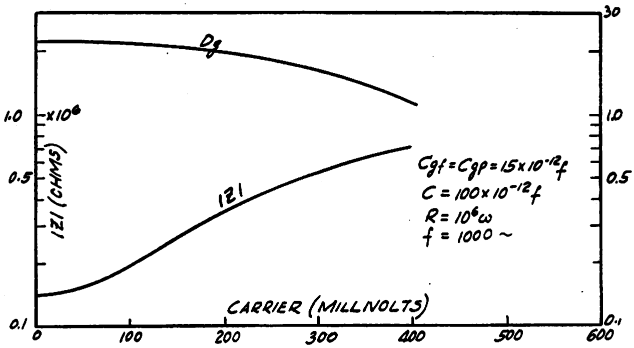

Fig. 4

A much more serious effect is found in the great increase of the r_p with the carrier voltage. The effect of this on the low audio frequencies is well known. In addition, it is found that for carrier voltages in the neighborhood of one volt and over, the great increase of r_p may actually cause a diminution of audible response. In connection with the selective properties of the radio-frequency amplifier this produces an interesting effect.

On adjusting the tuning condenser, approaching the setting for resonance with a strong modulated carrier, the loud speaker response (or the audio-frequency output of the detector) increases in accordance with the combined selective properties of the r-f amplifier and the rectifying property of the detector. The frequency distortion changes considerably all the while; at first the low frequencies are missing due to the selectivity of the amplifier. On approaching nearer the resonance setting the low frequencies begin to be heard and the output increases rapidly. At the same time, however, the r_p of the tube is increasing, and soon a setting is reached where the r_p is so great that the output of the detector diminishes, reaching a minimum at the setting of resonance. On passing over the resonance point the signal voltage on the detector decreases, the output increases and reaches a maximum, and then falls off again. A "double hump" has occurred, which has often unwittingly been attributed to causes in the radio-frequency amplifier.

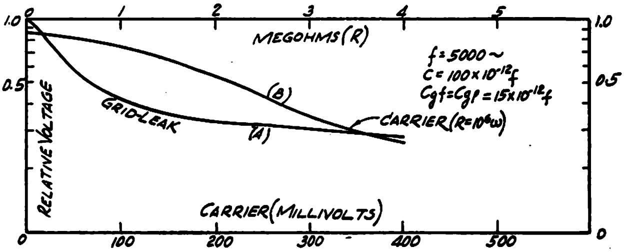

Fig. 5

A similar effect is found in the plate rectifier, although in this case it is due to the grid swing exceeding the bias voltage, so that the grid takes current. Up to the overload point, however, an increase of signal voltage causes a decrease of r_p improving the reproduction of the low audio frequencies. This is the opposite of what happens in the grid rectifier.

Fig. 6 shows curves obtained experimentally on a plate rectifier, for a weak signal, a signal just beginning to overload, and a signal

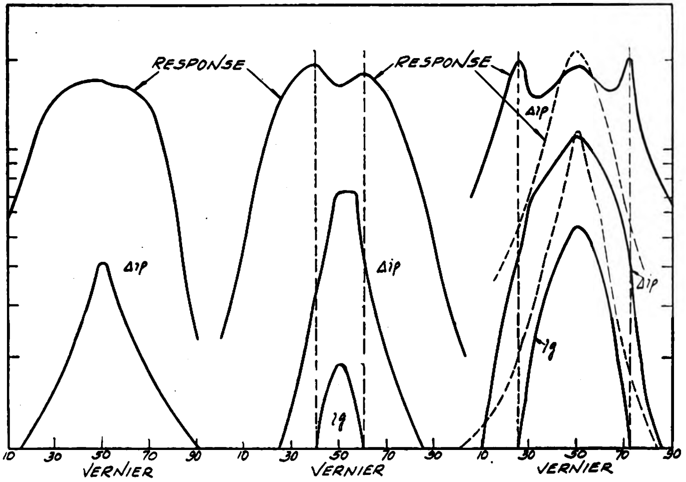

Fig. 6

which considerably overloads the detector. In the two latter cases grid current begins to flow exactly where the peaks occur. In the third case an additional peak occurs at the resonance setting, due to the fact that the increase of signal voltage exceeds the loss due to the flow of grid current. The effect of such overloading is to "chop off" the top of the response curve. Thus, in the case of the strongest signal, when the grid bias was increased to the point where the grid current just ceased to flow, and the plate voltage readjusted to furnish the same constant plate current, the broken curves were obtained. The apparent "broadness" of tuning due to detector overloading is evident.

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SINGLE-WIRE TRANSMISSION LINES FOR SHORT-WAVE ANTENNAS*

BY
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Summary—*The versatility of the possible arrangements of short-wave antennas can only be made available through the use of transmission lines. The phenomena of high-frequency transmission lines are discussed with respect to the effect of the low attenuation per wavelength and the influence of the termination on standing waves and radiation, for both an antenna and a transmission line.*

The single-wire transmission line is effective when properly terminated. It is the easiest method of feeding the Hertz antenna. The adjustment may be divided into two parts. The frequency which makes the Hertz antenna a pure resistance termination must first be determined. The proper point of connection between the line and antenna must then be found to make the terminating resistance equal to the characteristic impedance of the line.

When so terminated, experiment and theory show that radiation from the feeder will be small and that the feeder will act efficiently.

Theoretical and experimental curves to show the nature of a horizontal Hertz antenna are shown, since they influence the behavior of the line under changing conditions.

THE transmitting antenna is one of the most important and interesting of the component parts of a radio system. With the increased use of the shorter waves a greater flexibility is possible in the design of an antenna, and there is therefore a greater need for knowledge of its possibilities. When long waves are used, the shape of the antenna is generally determined by factors of cost, the largest item of which is the expenditure for the supporting towers. For instance, a vertical antenna to operate at a fundamental wavelength of 1000 meters would need to be about 750 ft. high, which would be very expensive. Therefore, long-wave antennas are usually of one general design, consisting of a flat top as high as can be built for the money available, with a vertical section connecting it to the transmitter at the bottom, the latter being coupled directly to the antenna circuit. At the shorter waves, however, the antenna is correspondingly smaller, so that it is possible to make it entirely vertical or horizontal, to operate it on harmonics, and in general to use a much greater latitude in design. One of the important advantages of this is that it makes it possible to construct the antenna so that the maximum radiation may occur at angles with the horizon such that the greatest effectiveness is obtained when this energy is reflected

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from the Heaviside layer. A great deal of work has been done on such arrangements, and this field of investigation appears very fruitful.

Since the limitations placed upon the antenna dimensions at longer wavelengths are removed in short-wave work, it often becomes desirable to separate the radiating antenna from the oscillator supplying the energy, for then all restrictions upon the placing of the antenna are removed. This necessitates the use of a transmission line to convey the energy from one to the other, without itself being a part of the radiating system. It is possible to design such a transmission line, but it is necessary to operate it properly if the desired result is to be obtained. It is the purpose of this discussion to outline what these proper conditions are.

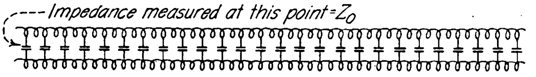
It should not be thought that the advantageous use of transmission lines is confined to short waves, as there are many fields of application in the medium and long-wave field. One of the most important is the possibility of removing the transmitter with its associated amplifiers from the strong field under the antenna, and by this means reduce the trouble due to feedback either in the radio-frequency amplifiers or in the audio amplifiers if there is some modulating action in the latter. This application becomes increasingly important as the power of the station is increased with a corresponding higher gain in the amplifiers.

GENERAL THEORY OF TRANSMISSION LINES

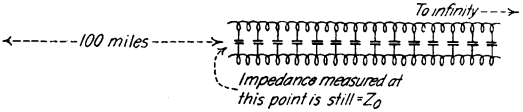
The action of transmission lines is a most absorbing study. It has been investigated thoroughly in its application to telephone and power. While the theory is quite general, some effects are more important in one field than the other. For instance, it is common to speak of the electrical length of the line, which is the length measured in terms of wavelength. In power use, the longest lines are about one-tenth of a wavelength long, the per cent power loss being comparatively small, while telephone lines may be several wavelengths long and the power received may be less than one per cent of that put into the sending end. In the radio transmission lines which we will discuss, their electrical length may also be several wavelengths, but the power loss will be small. It is apparent that it is desirable to give particular attention to a study of each application.

The fundamental treatment of a transmission line is usually mathematical in nature, but its physical concepts are quite easy to grasp and it is these concepts which are of importance to the engineer. Since they are often lost sight of in the formulas, consider physically what happens on an electrically long line, paying particular attention to the case where the energy dissipated along the line is small, as it is this case which has the most bearing upon lines operating at radio frequency.

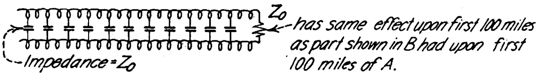
The outstanding physical fact about the transmission line is that its capacitance, inductance, leakance, and resistance are distributed evenly along the line instead of being in lumps, as is the case in the ordinary tuned circuit. The simplest sort of a line to consider is one which is infinitely long.


Impedance measured at this point = Z_0
To infinity

A
Infinite Line with Distributed Inductance Capacity and Resistance


To infinity
100 miles
Impedance measured at this point is still = Z_0

B
Infinite Line of A with 100 miles cut away


Z_0
has same effect upon first 100 miles as part shown in B had upon first 100 miles of A.
Impedance = Z_0

C
Line which cannot be distinguished from an infinite line by any measurements made along its length.

Fig. 1

If a line were infinitely long, and a length of say 100 miles is cut off, the line would still be infinitely long. Therefore, if the impedance of the line is measured before and after the piece is cut off, the same value will be obtained. This impedance is called the characteristic, surge, or iterative impedance of the line. Now if a study is made on the first hundred miles of the line, *there would be no way to distinguish between a line infinitely long and another one which was a hundred miles long and terminated in an impedance equal to that of an infinite line of the same characteristics.* This is illustrated in Fig. 1. We can therefore actually have a line which shows all the characteristics of an infinite line.

If an a-c generator were placed at the sending end of an infinite line, its effect would not be felt immediately everywhere along the line, but there would be a definite rate at which the disturbance would be propagated. This speed is not always the speed of light, but is generally less than that. On a loaded telephone cable, it is sometimes as low as 10,000 miles per second, but in the open wire line it is nearly that of light, and as the frequency increases and the magnetic flux within the wire decreases it approaches the speed of light as a limit, so that at radio frequencies the velocity is generally taken equal to approximately 300,000,000 meters a second.

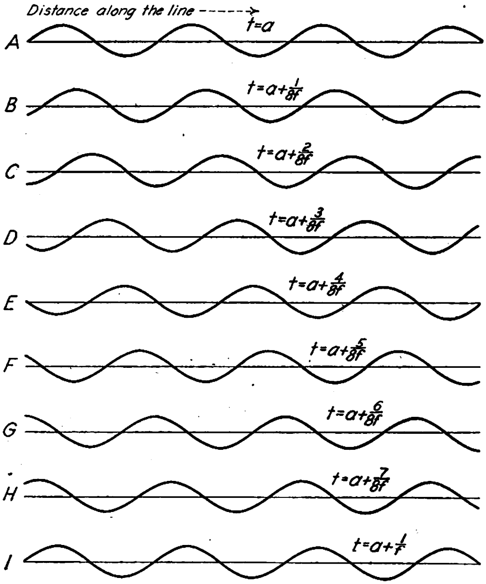


Fig. 2—Voltage distribution at different instants on line with negligible resistance.

If a group of oscillographs, or some other device which would indicate the instantaneous value of the voltage were connected at different points in the line, it would show that the voltage along the line varies sinusoidally with distance at any given instant. At some instant the distribution would be that shown in Fig. 2A. The distance between two maximum points in one direction is a wavelength and is equal to the velocity divided by the frequency. At a slightly later time the disturbance would have passed along the line and the voltage from point to point would be illustrated by Fig. 2B. And so as time passed the voltage would successively vary as shown in Figs. 2C to 2I. This cycle would be repeated as long as the voltage was maintained at the sending end. It is thus seen that there is a wave of voltage traveling down the line. In this figure it is assumed that the voltage is not attenuated as it passes along, or, in other words, that the losses due to resistance are negligible. In a telephone line the reduction in voltage due to resistance losses over a wavelength of line is generally quite large and the voltage at some one instant would be illustrated by Fig. 3, where the ratio by which the effective voltage is reduced by any given length of line is constant. This attenuation may be reduced by reducing the resistance or by increasing the inductive reactance of the line. When this is done by adding inductance to the line it is the process called loading. Since the inductive reactance is equal to $2\pi fL$, at high frequencies the inductive reactance is high, while the wavelength is short, making the resistance per wavelength small, so that the decrease in voltage over a few wavelengths is not very great. The current would also follow the same curves as the voltage and be in phase with it, for the characteristic impedance of an open wire line is practically a pure resistance at high frequencies, independent of frequency and equal to $\sqrt{L/C}$ where L and C are the values of inductance and capacity per unit length.

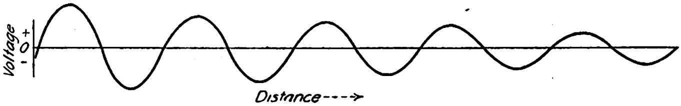


Fig. 3—Voltage distribution at some instant along line with appreciable resistance.

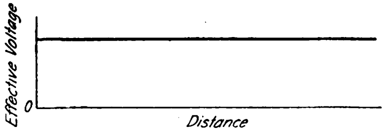


Fig. 4—Effective voltage as measured with a voltmeter along a line terminated in its characteristic impedance, showing that wave motion cannot be detected in this way.

If an ammeter and a voltmeter (which read effective rather than instantaneous values) were connected at various points along the line, since they read average effects, they would all have practically the same reading, and a plot of these values would be a straight line as shown in Fig. 4. It would not be possible with these instruments to detect the presence of the traveling waves.

The traveling waves have a counterpart in hydraulics, in the case of a canal of water very long and straight. If a disturbance is made at one end of the canal, waves will travel along, and a set of bobs on the surface would give the instantaneous values and draw curves similar to the instantaneous values of voltage and current shown in Fig. 2. If, however, the set of bobs is arranged so as to be moved up on rods in the water, but due to friction remain at the highest point of the wave after it has passed, so as to record maximum values, then a straight line would be obtained, similar to the values read by a voltmeter on the line as shown in Fig. 4.

Suppose instead of having the canal very long a board dam is created so that the waves will strike it. There will then be a piling up of water at the dam which will start a wave back in the opposite direction.

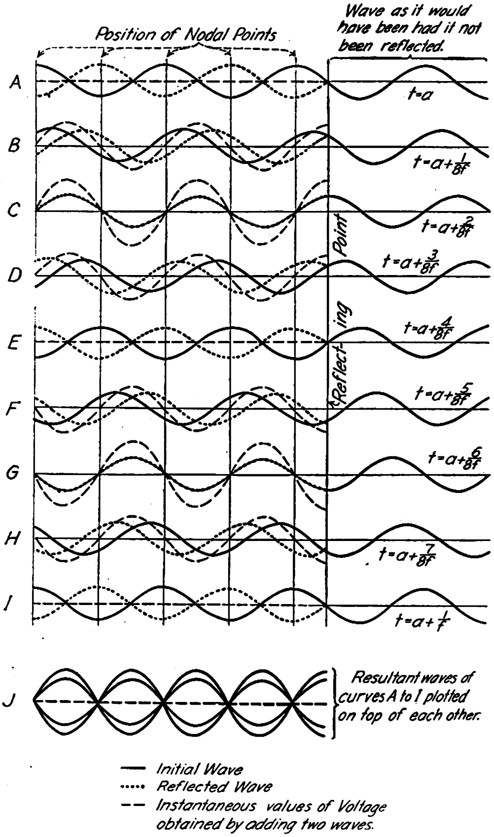

Position of Nodal Points
Wave as it would have been had it not been reflected.
A
B
C
D
E
F
G
H
I
J
$t=a$
$t=a+\frac{1}{8f}$
$t=a+\frac{2}{8f}$
$t=a+\frac{3}{8f}$
$t=a+\frac{4}{8f}$
$t=a+\frac{5}{8f}$
$t=a+\frac{6}{8f}$
$t=a+\frac{7}{8f}$
$t=a+\frac{1}{f}$
Reflecting Point
Resultant waves of curves A to I plotted on top of each other.
— Initial Wave
····· Reflected Wave
— — Instantaneous values of Voltage obtained by adding two waves.

Fig. 5—Distribution of voltage on open circuited line.

In the same way, if the electric line is not infinitely long or terminated in its characteristic impedance, but instead is open circuited, then when the wave strikes the end, it also will be reflected. The electric wave in traveling has half of its energy stored in the magnetic field due to the current, and half of the energy is stored in the electric field due to the voltage. When it strikes the open circuit, the magnetic field will collapse, for the current must become zero. But a changing magnetic field produces an electric field. It is upon this principle that all generators and transformers operate. The energy stored in the magnetic field will therefore be turned into energy in an electric field and added on to the existing field, so that the voltage at the end circuit will be increased. This increased voltage will start a wave traveling back in the opposite direction, and since there has been nothing to absorb any energy at the open circuit, the returning wave will be of the same magnitude as the original wave. As the electric field starts moving back, it will set up a magnetic field again, and the energy will once more be divided equally between the two. As the electric field has simply been doubled at the instant of reflection, the voltage of the returning wave starts out in the same phase as the original wave, but the magnetic wave, and hence the current of the reflected wave, is in opposite phase to the incident wave at the open circuited point. The total voltage and current at any point and any instant is therefore the sum of the voltages or currents of the incident and reflected waves. That the current has been reversed in phase is evident from the fact that the two current waves must add to zero at the point of open circuit, and since the two voltages are equal and do not add to zero, they must remain in phase.

In Fig. 5A is shown a plot of the two voltage waves which are present on the open circuited line at some instant. The total voltage at any point on the line at that instant is the algebraic sum of the two waves, and is also shown. It will be noticed that the reflected wave is obtained by folding back on the line from the point of reflection, the initial wave as it would have been beyond the open circuit had the line continued on. In Figs. 5B to 5I this process is continued for later instants. When the two waves are added point by point it will be seen that at some points the resultant wave is always zero. These points occur at odd quarter wavelengths from the open circuit, because at such a position the reflected wave has traveled an odd number of half wavelengths since it passed this point as part of the initial wave. At even quarter wavelengths from the open circuit, the resultant voltage is always twice the value of either wave, since the two waves at these points are equal and in the same direction. In Fig. 5J all the resultant waves obtained for the different instants are plotted on top of each other to permit a better comparison. It will be noticed that voltage varies with time and space. At every point along the line the voltage reaches a maximum at the same instant, but this maximum voltage varies sinusoidally along the line. It should be noticed in contrast, that on the line terminated in its characteristic impedance the maximum voltage occurs at different instants for different points on the line but is the same value for all points. Since the effective voltage is the maximum divided by the $\sqrt{2}$, if a voltmeter is moved along the open circuited line, a curve similar to Fig. 6 would be obtained, the readings being always positive since the meter cannot take account of phase. Waves which can be detected by a meter which records average results are called standing waves, and are different from the traveling waves which were first discussed. This is apparent after comparing Figs. 4 and 6. The points of minimum voltage are called the nodes of voltage, while the points of maximum voltage are called the antinodes. The voltage at the nodes is not actually zero since there is always some loss due to resistance as the wave travels along the line, and the returned wave being slightly smaller will not entirely neutralize the initial wave at these points, but at high frequencies there is a close approximation to this ideal condition.

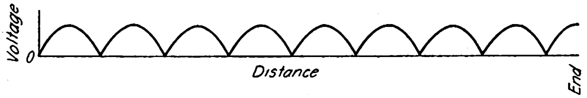


Fig. 6—Voltage readings along an open circuited line with negligible resistance showing standing waves. (Voltmeter cannot record phase so all readings are positive.)

Figs. 7A to 7I show the two waves of current on the line for different instants. It should be remembered that while a voltage wave and a current wave are discussed as though they were distinct, the voltage and current together are necessary for a physical wave, since one cannot exist without the other, but each can be considered separately in plotting results. In the figure it will be noticed that the current has been reversed at the open circuit, so that the two waves add up to zero at that point. As the instantaneous values are added point by point it will be observed that there is also a standing wave of current. This wave is the same as would be obtained for voltage if the reflection had occurred a quarter wavelength down the line, and the nodes of current occur at the antinodes of voltage and vice versa.

In any line the numerical ratio and phase between voltage and current at any point in the line is a measure of the impedance beyond that point. In the line terminated in its characteristic impedance the ratio is the same at every point and the voltage and current are in phase, and any length of line therefore has no natural period. In an open circuited line this ratio varies along the line, being small at the nodes of voltage and large at the nodes of current. By examining this ratio, the way the impedance of an open circuited line varies can be determined. In Fig. 8 the effective values of voltage and current are plotted on top of each other. Immediately below is plotted their ratio. This shows how the impedance varies as the distance from the open circuited end is varied, and is therefore a plot of impedance vs. length of line. This is also the way the reactance of an antenna varies since it is a line with distributed constants. It will be noticed that the line is resonant at odd quarter wavelengths, and antiresonant at even quarter wavelengths. Since the effective voltage at any point on the line is represented by the equation $E_x = E_m \cos(2\pi x/\lambda)$ and the current is represented by a similar equation $I_x = I_m \sin(2\pi x/\lambda)$ their ratio is represented by the equation

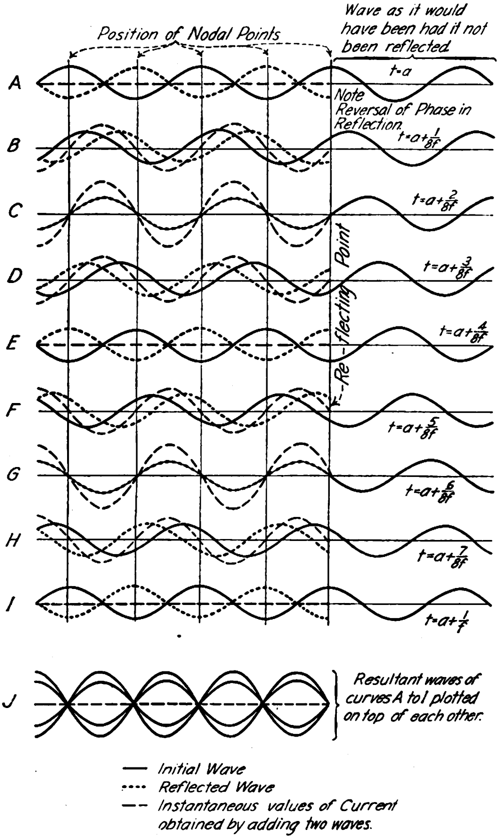

Position of Nodal Points
Wave as it would have been had it not been reflected
A
t=a
Note Reversal of Phase in Reflection
B
t=a+1/8f
C
t=a+2/8f
D
t=a+3/8f
E
t=a+4/8f
F
t=a+5/8f
G
t=a+6/8f
H
t=a+7/8f
I
t=a+1/f
Re-flecting Point
J
Resultant waves of curves A to I plotted on top of each other.
Initial Wave
Reflected Wave
Instantaneous values of Current obtained by adding two waves.

Fig. 7—Distribution of current on open circuited line.

$$Z_x = \frac{E_x}{I_x} = \frac{E_m}{I_m}\,\frac{\cos\dfrac{2\pi x}{\lambda}}{\sin\dfrac{2\pi x}{\lambda}} = Z_0 \cot\frac{2\pi x}{\lambda}$$

In these equations the distance x is measured from the open end of the line, the subscript x is used to denote the values at the distance x from the end, and the subscript m is used to denote maximum values (those that occur at the antinodes).

It is not enough ordinarily to know how the absolute value of an impedance varies, but it is also desirable to know whether it is a resistance, capacitive reactance, inductive reactance, or an impedance with both resistance and reactance. This can best be determined by means of vector diagrams.

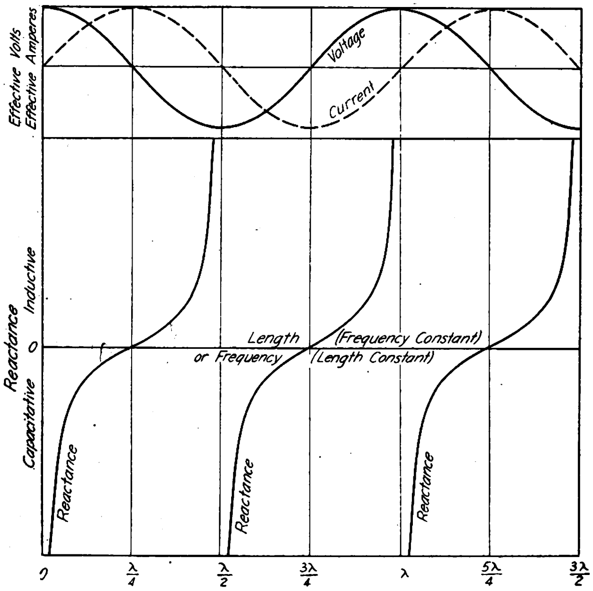
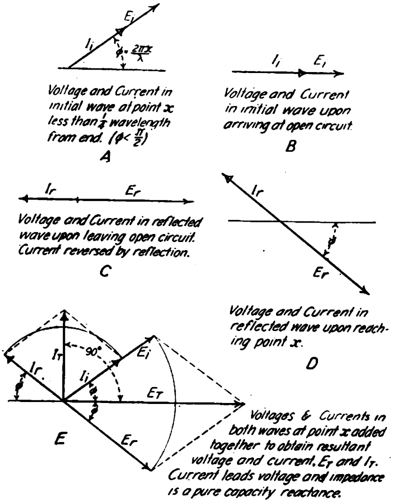


Fig. 8—Variation of impedance, open circuited line.

Fig. 9 shows a progressive vector diagram for a point less than a quarter wavelength from the end. Fig. 9A shows a vector representing the voltage and current of the initial wave at the point in question. The voltage and current of the initial wave at the open circuit are shown in Fig. 9B and lag the vectors in Fig. 9A by an angle $\phi = 2\pi x/\lambda$. Upon reflection the reflected wave voltage and current will be represented by Fig. 9C. The fact that these are 180 deg. out of phase indicates that energy is now being propagated in the opposite direction. After traveling back along the line, the vectors at the point x are shown in Fig. 9D and lag the vectors in Fig. 9C by a similar angle ϕ. In Fig. 9E the two sets of vectors representing the two waves are added together and give the resultant voltage and current E_T and I_T. It will be noticed that if the attenuation is negligible, so that the currents and voltages in the two waves are equal, the resultant voltage and current will be at right angles to each other, and in this case when ϕ is less than 90 deg. the current leads the voltage and the impedance of such a length of line is a pure capacitive reactance.

In a similar manner Fig. 10 shows the vector diagram at a point between one quarter and a half wavelength from the open end. In this case the current and voltages are also 90 deg. apart, but the current lags the voltage and the impedance is therefore a pure inductive reactance.



Voltage and Current in initial wave at point x less than $\frac{1}{4}$ wavelength from end. ($\phi < \frac{\pi}{2}$)
A

Voltage and Current in initial wave upon arriving at open circuit.
B

Voltage and Current in reflected wave upon leaving open circuit. Current reversed by reflection.
C

Voltage and Current in reflected wave upon reaching point x.
D

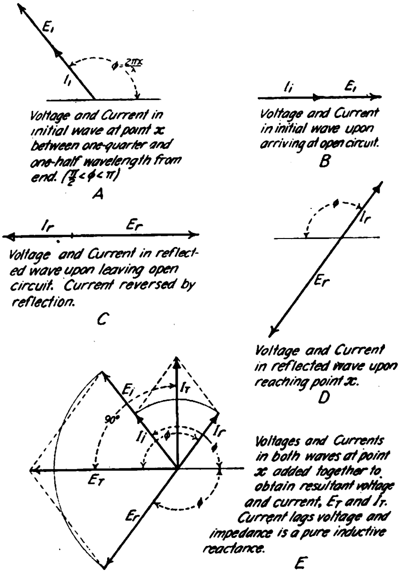
Voltages & Currents in both waves at point x added together to obtain resultant voltage and current, E_T and I_T. Current leads voltage and impedance is a pure capacity reactance.
E

Fig. 9—Progressive vector diagram of voltage and current at point on line less than one-fourth wavelength from end. Attenuation of line assumed negligible.

Fig. 11 shows the result of attenuation upon the impedance of the open line. It is drawn for a length less than one quarter wavelength, but a similar diagram can readily be constructed for any length. Due to attenuation from resistance losses the voltage and current of the returned wave are less than those of the initial wave. The resultant sums of voltage and current are now less than 90 deg. apart, showing that the impedance has a resistive component.

These results are to be expected from the fact that if the line has no loss, all the energy which goes into any length is later returned, a condition similar to that produced by a pure reactance. On the other hand, if the line does dissipate energy, all the energy will not be returned and the impedance of any length will have a resistive component.

If instead of varying the physical length of the line, it is kept constant and the frequency is varied, the *electrical* length of the line will be varied just as effectively as when the frequency was constant and the physical length of the line changed. The impedance curve of Fig. 8 might therefore have frequency for its abscissa instead of distance. It is possible, therefore, to obtain any reactance desired by varying either the length or frequency.



Voltage and Current in initial wave at point x between one-quarter and one-half wavelength from end. ($\frac{\pi}{2} < \phi < \pi$)
A

Voltage and Current in initial wave upon arriving at open circuit.
B

Voltage and Current in reflected wave upon leaving open circuit. Current reversed by reflection.
C

Voltage and Current in reflected wave upon reaching point x.
D

Voltages and Currents in both waves at point x added together to obtain resultant voltage and current, E_T and I_T. Current lags voltage and impedance is a pure inductive reactance.
E

Fig. 10—Progressive vector diagram of voltage and current at a point on line between one-fourth and one-half wavelength from end. Attenuation of line assumed negligible.

So far only two terminations for a transmission line have been discussed, viz., termination in the characteristic impedance or in an open circuit. From the relations deduced for the open circuit it is possible to see what the result would be when the line is terminated in a pure reactance. Since there would not be any absorption of energy in the termination, the returned wave would be as strong as the initial wave, and therefore the same sort of standing wave would result. This may also be seen because, in passing back from the open end of a line, the short length out to the end may be made a pure reactance of any value desired including zero or a short circuit. Such a short length could be replaced by a pure reactance of the same value, and would give the same results as a termination for the rest of the line. It can therefore be seen that terminating the line in a pure reactance of any value whatever will merely shift the standing wave along the line, but will not change its character.

If, in the hydraulic model mentioned earlier in the article, instead of having a dam so as to reflect the wave completely, a submerged board had been placed in the canal, then some of the energy would have been reflected and some would have gone on. The returning wave would therefore not be as strong as the original wave.

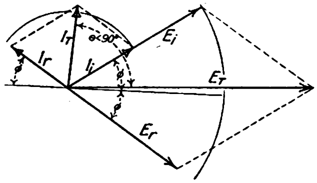

I_T
$\theta < 90°$
E_i
I_r
I_i
ϕ
E_T
E_r

Fig. 11—Vector diagram of voltage and current at point on line less than one quarter wavelength from end. Attenuation makes impedance partly resistive.

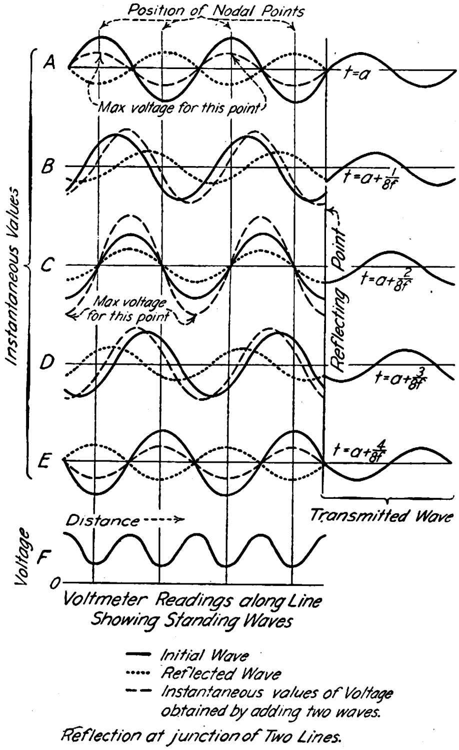

Position of Nodal Points
A
$t=a$
Max voltage for this point
B
$t=a+\frac{1}{8f}$
C
$t=a+\frac{2}{8f}$
Max voltage for this point
D
$t=a+\frac{3}{8f}$
E
$t=a+\frac{4}{8f}$
Instantaneous Values
Reflecting Point
Transmitted Wave
Distance
Voltage
F
0
Voltmeter Readings along Line Showing Standing Waves
Initial Wave
Reflected Wave
Instantaneous values of Voltage obtained by adding two waves.
Reflection at junction of Two Lines.

Fig. 12—Distribution of voltage on line when terminated in an impedance greater than the characteristic impedance.

If two transmission lines of different characteristics are joined together, or if the single line is terminated in something other than the characteristic impedance, reflection will also occur, with a weaker returning than incident wave. Standing waves will again be produced, but since the two waves can never cancel each other at any point, the current and voltage will never be reduced to zero, but only to a minimum value. Figs. 12A to 12E show a plot at different instants of instantaneous values of voltage for the case of partial reflection, where the return wave is half the magnitude of the original wave. Fig. 12F shows the plot of effective values of the voltage along the line and is obtained by examining curves *A* to *E* and determining the maximum voltage which ever occurs at each point of the line. If the impedance of the termination is greater than the characteristic impedance, the minimum points of current and voltage will occur at the same points as the nodes of the open circuited line, while if the terminating impedance is less than the characteristic impedance the minimum points will correspond to the nodes of a short circuited line. In both cases the minimum points of voltage will be maximum points of current and vice versa.

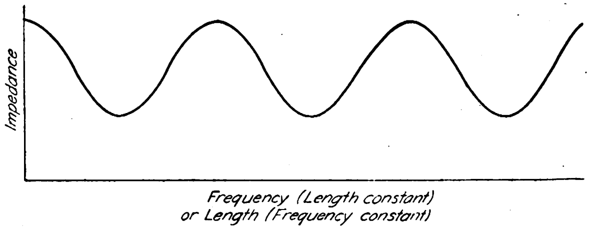


Fig. 13—Variation of impedance of line terminated in impedance greater than characteristic impedance.

In the same way the impedance will not be a pure reactance but will vary in magnitude between maximum and minimum values in the manner shown in Fig. 13. Here again the abscissa may be length for a fixed frequency, or frequency for a fixed length. This variation in impedance with frequency is used to locate faults which develop in long telephone lines, and is a measurement of the time it takes the waves to pass down to the fault and return after being reflected. It therefore corresponds in principle to recently developed methods of sounding the sea by measuring the time taken for a sound to pass to the bottom and return after reflection. A similar principle is used to detect pockets in earth oil fields before undergoing the expense of sinking a well.

That this case of reflection is a definite and not a theoretical phenomena is illustrated in a very real trouble which is encountered due to echoes from impedance irregularities on loaded telephone lines where the velocity of propagation is low. Since telephone repeaters amplify in both directions, the reflected wave is amplified along its return path, and may differ sufficiently in time and strength from the original wave to interfere with speech.

Usually reflection of a wave decreases the efficiency of a line, as the wave in passing back along the line causes additional losses. There are some exceptions in telephone work where the characteristic impedance of a cable may have a considerable capacitive component, when the best termination from the standpoint of efficiency is an impedance equal in magnitude to the characteristic impedance of the cable but having an inductive instead of a capacitive reactance. Also if the generator is not subject to adjustment and the maximum power is to be obtained from the generator, another termination may at times be desirable on shorter lines. However, for high-frequency work, the characteristic impedance of the transmission line, as has been mentioned before, is practically always a pure resistance, and since adjustments may be made at the generator, the termination desired is that which will produce no reflection.

To summarize, it can be seen that:

(a) A line terminated in its characteristic impedance has the same characteristic as an infinite line.

(b) A line terminated in its characteristic impedance will have no standing waves, the traveling waves producing the same effect on a voltmeter or ammeter anywhere along the line.

(c) A line terminated in its characteristic impedance will have a constant impedance at high frequencies which is equal to a pure resistance, and therefore has no natural period.

(d) A line terminated in an open or short circuit will have a standing wave set up on it, and the maximum points of voltage will be the minimum points of current and vice versa.

(e) A line terminated in an open circuit will be a pure reactance if the resistance is negligible, or if there is resistance, will be largely reactive except at the fundamental and odd harmonics when it will be a low resistance. The reactance will be alternately inductive and capacitive if either frequency or length is varied.

(f) A line terminated in anything but the characteristic impedance will have standing waves set up on it, but the variation between the maximum and minimum points of current and voltage will decrease as the value of the terminating impedance is made to approach that of the characteristic impedance.

(g) A line terminated in anything but the characteristic impedance will vary in impedance with frequency or length, and will therefore have a natural period, but the magnitude of this variation in impedance with frequency will decrease as the terminating impedance is made to approach the characteristic impedance.

(h) A line for high-frequency work is most efficient when terminated in its characteristic impedance.

Thus the fundamentals are set forth upon which the proper design of an antenna and transmission line feeder can be determined.

TRANSMISSION LINES TO FEED ANTENNAS

One of the most versatile types of antenna is the so-called Hertz antenna, consisting of a single wire, without a counterpoise, in which the voltages at the two ends of the wire are opposite in phase. Its versatility lies in the fact that by operating it horizontally, vertically, or at some other angle to the earth, by varying its height above the earth, and by operating it either at its fundamental, or one of the harmonics, a wide variety of radiation patterns may be obtained. To take advantage of this versatility, and to predetermine its action, some type of transmission line feed is necessary to convey the energy from the driver to the antenna. Two types of transmission lines are in use, the two wire and the single wire feed. These have been called respectively "current" and "voltage" feeds, but it is not believed that this is a very good term, as it conveys an erroneous impression of their action. In each case, as in any case where power is transmitted, both current and voltage are necessary, and contrary to the opinion entertained when this investigation was started, it has been found that the characteristic impedances and hence the voltages and currents for a given amount of power do not differ very much between the two types of lines.

In order to minimize radiation from the feeder, it must be terminated in its characteristic impedance and thus prevent the formation of standing waves. The radiation resistance of an efficient antenna is lower than the characteristic impedance of any feasible transmission line and so means must be used to match the one against the other. In the two-wire line this may be done by means of a network equivalent to a transformer, designed for the particular frequency to be used and inserted between the line and the antenna. It is believed, however, that the single-wire transmission line has a number of advantages, which usually make it more desirable. One of the most important of these is that it is readily possible to terminate it properly by the antenna without the use of a transformer. It is more convenient to construct, especially if the antenna is suspended at some height above the ground, where the proper spacing of the two-wire feed would be awkward to secure. It is very easy to adjust when the wavelength is changed or when the radiation of the antenna is changed by changing its height. It is possible to change the antenna system readily from operation on the fundamental to operation on some harmonic. It should be noticed that while a grounded antenna, or a Hertz with the driver connected permanently in the center, can only be operated on odd harmonics of the fundamental, the use of a transmission line makes it possible to operate a Hertz antenna on both odd and even harmonics. It has even been found with the single-wire feed that by selecting the proper height of antenna to adjust the radiation resistance, a single physical construction could be secured in which the single-wire feeder was properly terminated for both the fundamental and the second harmonic, affording a ready means of shifting from one wave band to another. These advantages are particularly important in making investigations where it is desirable to change the construction of the antenna, or the operating wavelength frequently.

One factor which has brought the single-wire line into disrepute is that, when improperly adjusted, it will radiate much more than the two-wire line, since the latter has an action similar to a loop, in which the radiation is a function of the separation of the two sides, measured in wavelengths. By making this separation small, the radiated power of an improperly terminated two-wire line may be quite small. When the proper adjustment of the single-wire line is made, its operation will be satisfactory, and it is the purpose of this discussion to indicate how those adjustments should be made.

Carson[1] has shown in a digest of the literature that radiation is negligible on a two-wire line when terminated in its characteristic impedance, and the same point was brought out by Mannebach.[2] In the discussion of the latter's paper, Slepian[3] gives a good description of the mechanism of radiation due to a principle of relativity which applies with equal force to the single-wire line. In this description the radiation occurs only when there is a snapping of the lines of force due to reflection at the end of the line. This occurs in all antenna systems.

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[1] J. R. Carson, "Radiation from Transmission Lines," *Jour.* A. I. E. E., pp. 789–790, 1921.

[2] Charles Mannebach, "Radiation from Transmission Lines," *Jour.* A.I.E.E., pp. 95–105, 1923.

[3] Slepian, Discussion, *Jour.* A. I. E. E., pp. 981–982, 1923.

When a single electron is moving at a constant velocity, no radiation occurs. If the electron is accelerated or decelerated, radiation will take place. However, the radiation is due, not to the acceleration or deceleration of the electron itself but to the acceleration of a portion of the field, although it is realized that the two cannot be separated in the case of the single charge.

Similarly if a series of doublets along a transmission line are so arranged in phase that the field is propagated along the line with a constant velocity, no radiation will take place. This constant velocity occurs when there is a continuous shift in phase of the doublets, so that even though the electrons flow first in one direction and then the other, the resultant field moves with constant speed. This condition is produced on a transmission line when there is no reflection and therefore no returning wave.

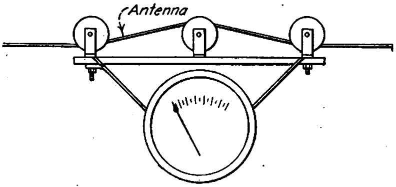


Fig. 14—Trolley for obtaining current distribution along antenna.

By application of the reciprocity theorem and a study of wave antennas it will be concluded that there will be a small radiation in the direction of the transmission line.

Hence this theory is postulated: Radiation is largely due to reflection.

In the previous discussion it has been shown that reflection always produces standing waves, and consequently the test for the absence of radiation on a transmission line is the absence of standing waves.

Actual measurements on the line show a high efficiency for the transmission line and therefore experimentally verify the absence of appreciable radiation.

A single-wire feeder was connected to a Hertz antenna at some point away from the center of the antenna, and current distributions over the system for three different frequencies were measured by the trolley arrangement shown in Fig. 14 in order to verify the theory. This trolley was used to shunt a meter across a portion of the antenna about eight inches long. The current in the meter is therefore proportional to the total current and the combination may be calibrated and the actual current known. By means of a system of cords and pulleys the trolley was drawn along the antenna or transmission line and the distribution of current determined. The meter readings were made by means of a transit. A tube working below normal plate voltage was used in order that fluctuations due to overheating of the tube during steady operation might not be encountered.

Curve A, Fig. 15, shows the distribution for a wavelength greater than the fundamental. Curve B shows the distribution for a wavelength less than the fundamental. Curve C shows the distribution when the wavelength is equal to the fundamental of the antenna. There are two interesting features here, first that at the fundamental wavelength the currents on the two sides of the feeder are equal, and secondly, that the current at the center of the antenna may be greater when the operation is at a higher wave than the fundamental. However, while in this case the current at the center is greater,the current at the short end is smaller, and the total radiation is less than when the antenna is operated at its fundamental. This definitely shows that such an antenna system cannot be adjusted to its fundamental by placing a meter in the center and adjusting for maximum current. The criterion which was adopted was to place a meter on either side of the feeder and adjust until the two currents are equal. The intersection of the two curves of current will be the fundamental wavelength of the antenna.

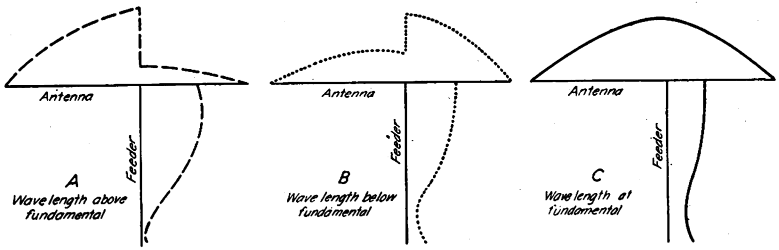

Antenna
Feeder
A
Wave length above fundamental
Antenna
Feeder
B
Wave length below fundamental
Antenna
Feeder
C
Wave length at fundamental

Fig. 15—Distribution of current on antenna and feeder for three different wavelengths (feeder tap not yet adjusted).

To test this criterion and to show the independence of the fundamental wavelength from the position of the feeder, curves were taken for different positions of the latter. These are shown in Fig. 16. It will be noticed that the intersections of these curves all occur at the same wavelength.

It can readily be understood why the currents will vary in the manner shown in Fig. 16. Refer to Fig. 8 which shows how the reactance of a given length of line varies with frequency. At the junction of the feeder and antenna there are two branches of the antenna. The short end will have a capacitive reactance and the long end will have an inductive reactance, but at the fundamental wavelength, these two reactances will be equal and the currents into the two branches will be equal. If now the length of the longer one is increased and the length of the shorter branch is reduced by the same amount (which corresponds to moving the feeder tap) the reactance of both branches will increase by equal amounts as is apparent from the symmetry of Fig. 8. Therefore, the two currents will remain equal to each other as the feeder is moved outward.

If now the feeder is in a given position and the frequency is increased, the reactance of the short end will drop, while the reactance of the long end will rise and the currents in the two branches will be unequal. Since the distribution of current along the antenna is sinusoidal in space and the wavelength has been decreased by the rise in frequency, the two sine waves will not have equal magnitudes at the junction of the feeder, and the current distribution will be as shown in Fig. 15B.

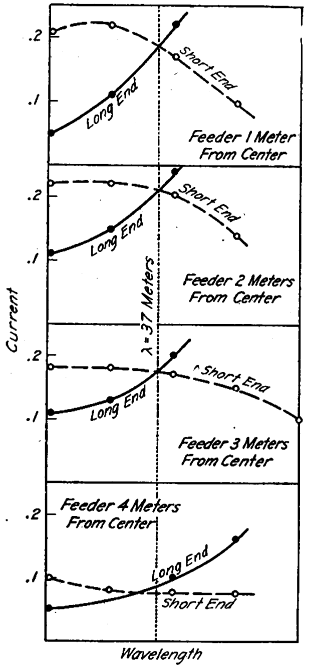
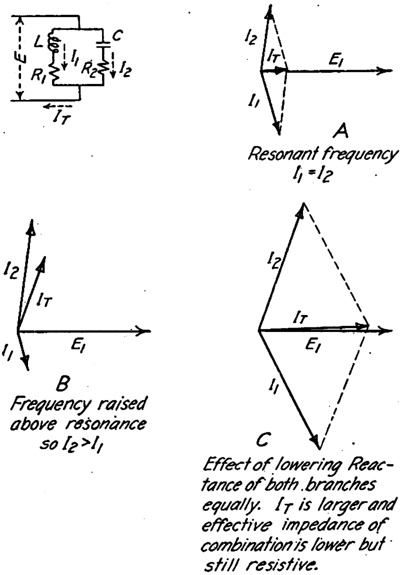


Fig. 16—Variation of current in two branches of antenna as wavelength is varied

In a similar manner if the frequency is lowered, the reactance of the two branches would change in opposite directions, and the two currents will be unequal. As the wavelength has now been increased the current distribution in Fig. 15A will be obtained.

The two branches of the antenna correspond to two branches of a parallel circuit, one of capacitive reactance and the other of inductive reactance. Each branch has resistance due to the ohmic and radiation resistance of the antenna. If in such a parallel circuit the frequency is adjusted until the currents in the two branches are equal, the sum of the two currents will be less than either, and will be in phase with the voltage. This is shown by the vector diagram of Fig. 17A. The impedance of this combination is therefore a pure resistance, or very nearly so. If, however, the frequency is such that the currents are not equal, then the sum of the two currents will be out of phase with the voltage and the impedance will be largely reactive, provided the reactance of the two branches is large in comparison with the resistance. This is shown in the vector diagram of Fig. 17B. If in the parallel circuit, with the frequency adjusted to make the two currents equal, the reactances of both branches are decreased, but still are kept equal to each other, and the total resistance is not changed, then the two currents will increase for the same voltage, and the total current will increase. The effective impedance of the combination will be nearly a pure resistance, but of a lower value. This is shown by the vector diagram of Fig. 15C. The resistance of a parallel circuit tuned to resonance is approximately X^2/R_1+R_2,



Resonant frequency $I_1=I_2$

Frequency raised above resonance so $I_2>I_1$

Effect of lowering Reactance of both branches equally. I_T is larger and effective impedance of combination is lower but still resistive.

Fig. 17

where X is the reactance of either branch, R_1 and R_2 are the resistances of the two branches, respectively. As the point of connection of the feeder is moved along the antenna, the reactance of the two branches is simultaneously varied, being reduced as the center of the antenna is approached. The approximate formula for the variation of input resistance was derived and is

$$R_i = \frac{Z_0^2}{R_r} \cos^2 \frac{\pi x}{l}$$

where

R_i = effective input resistance
R_r = effective resistance of antenna measured at center
x = distance of connection from the end
l = length of antenna.

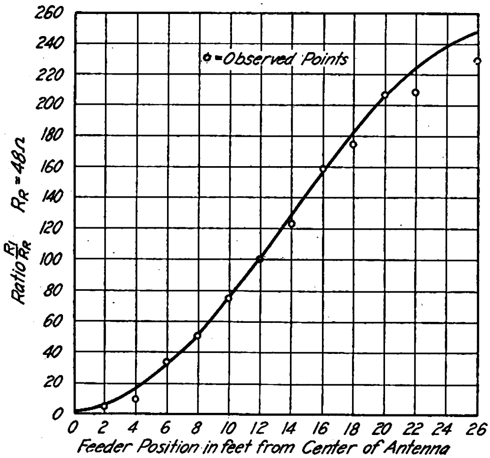


Fig. 18—Theoretical input resistance for various feeder positions.

A more accurate formula is the following:

$$R_i = \frac{R_r x(l-x)}{2 \sin^2 \pi l} + \frac{Z_0^2}{R_r} \cos^2 \frac{\pi x}{l}.$$

The first term of this equation is of importance only near the center of the antenna. To test the formula the input power was measured. This divided by the square of the current in the feeder at the junction gave the input resistance of the antenna and its effective resistance as a termination for the transmission line. Fig. 18 shows a plot of the results obtained, which is nearly a cosine squared curve, with a plot of the theoretical curve, which deviates appreciably only at the outer end of the antenna. This deviation is to be expected since at the outer end the reflection on the feeder is pronounced, and it is therefore radiating a considerable amount of power, which will also produce an effect upon the radiation resistance of the antenna.

It is possible to obtain a proper termination for the transmission line. By adjusting the frequency to the natural period of the antenna, the termination of the line can be made pure resistance. This is a first essential, as it has been pointed out that the characteristic impedance of the transmission line is a pure resistance. Any other frequency will not work (except harmonics of the fundamental), for then the termination would be largely reactive, and no matter what the value of this reactance, reflection will occur and standing waves will be set up on the line.

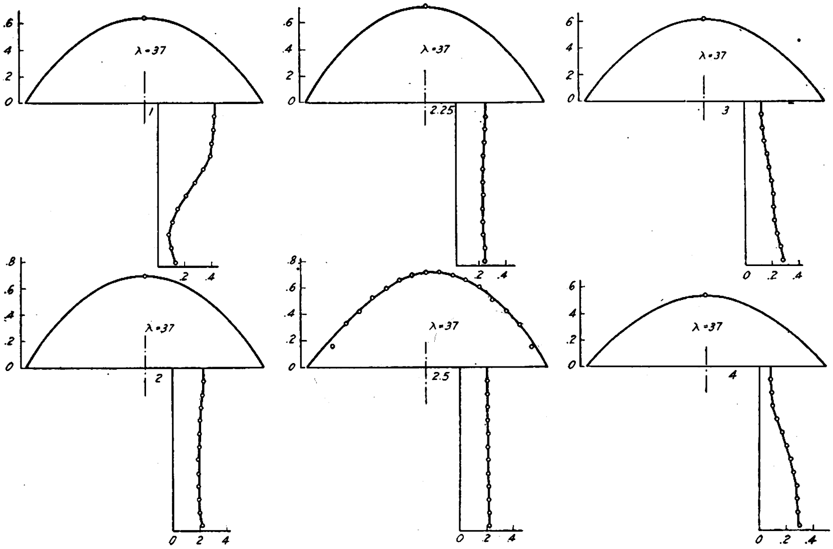


Fig. 19—Distribution of current on feeder and antenna for different positions of feeder.

The job is not quite finished when the proper frequency has been selected to make the termination a pure resistance, for the resistance must be adjusted to the proper value, which can be done by moving the tap out along the antenna until standing waves are eliminated upon the feeder. Fig. 19 shows the distribution of current for taps at different distances from the center. In this case the antenna and feeder were No. 12 wire, and the antenna was fifty feet high and eighteen meters long.

If the frequency is adjusted to give maximum current in the center of the antenna, a wavelength longer than the fundamental will always be obtained. This gives a very violent reflection at the junction, and sets up large standing waves on the feeder, and the system becomes nothing but a T antenna.

The standing waves obtained when the right frequency is used but the tap is not properly adjusted were never quite so violent. That this adjustment is important, however, is shown by the curve of Fig. 20, where with a constant adjustment at the generator the current in the center of the antenna varied with the position of the feeder tap and the maximum current was obtained at the point where standing waves were eliminated. It should be noticed that while it is not a good criterion to obtain the maximum current at the center when the frequency is varied, since the distribution will be altered, yet maximum current is a good criterion when anything else is varied, since the relative current distribution on the antenna will remain constant if the frequency is fixed. The same antenna was operated at its second harmonic and the distribution of current is shown in Fig. 21.

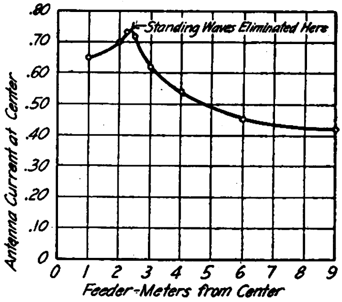


Fig. 20—Curve showing maximum antenna current at point where standing waves in feeder are eliminated. Antenna at its fundamental.

While the method described for finding the fundamental frequency is easy to apply, since it involves only connecting a meter to either side of the junction of antenna and feeder, the method given for determining the proper terminal point is more tedious, requiring as it does, a means of measuring the current along the line. Since reflection will tend to make the input impedance of the line reactive, any device which will determine when the line is at unity power factor may be used if the feeder is not an integral number of quarter wavelengths long.

This determination will be most accurate if the transmission line is an odd number of eighth wavelengths long, as then the power factor will change to the greatest extent when reflection is present. One method of determining when the power factor is unity, which has been tested and found to be fairly accurate, is to connect between the feeder and ground a meter through a condenser, and connecting on either side of the first meter another meter. The power factor may then be determined by the three-meter method. Another method which gives approximate results with a Hartley oscillator is to clip the feeder on the driver circuit. If the feeder and system is inductive it will tend to reduce the wavelength of the oscillator; if it is capacitive, it will tend to raise it, while if it is resistive, it will not ordinarily have much effect upon the frequency.

It might appear that the proper position of the junction of the feeder and an antenna could be specified in fractions of a wavelength. This cannot readily be done because of the oscillating character of the antenna constants as the height is increased. By the method of Abraham[4] and Pierce[5] the expression for the radiation resistance of a horizontal Hertz antenna operating at its fundamental was determined to be

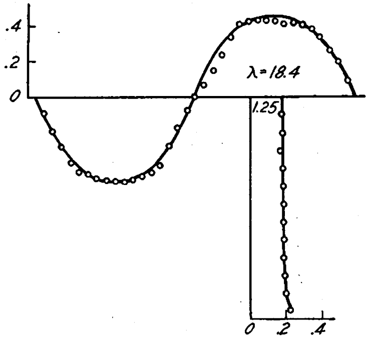


Fig. 21—Antenna working at second harmonic with correct feeder termination

$$R_R = \frac{480}{\pi} \int_0^{\pi/2} \int_0^{\pi/2} \frac{\sin^2 (\pi h/l \sin\theta \sin\phi) \cos^2 (\pi/2 \cos\theta)}{\sin\theta} d\theta d\phi.$$

By graphical integration this expression was evaluated and the results obtained are shown in Fig. 22, where the abscissa is the ratio of height to length. Pierce and Abraham obtained a radiation resistance of a vertical quarter wave-antenna of 37 ohms, so that a half wave-antenna removed from the earth should have a resistance of 74 ohms. It will be seen that the computed values approach this in an oscillating manner. In this respect they are similar to the curves obtained by Levin and Young,[6] in their analysis of the vertical antenna at different heights above the ground.

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[4] M. Abraham, *Phys. Zeits.*, **2**, 329–334, 1901.
[5] G. W. Pierce, "Electric Oscillations and Electric Waves," 1920.

The curve was checked by measurement, and the results shown in Fig. 22 were obtained. In the theoretical formula the earth is considered as a perfect reflector. In Fig. 22, it is apparent that if the ground plane had been assumed as two meters below the ground, the curves would nearly coincide.

A test with a very long feeder, mentioned later, shows experimentally that the radiation of a properly terminated feeder is small. The effective resistance of the antenna may therefore be conveniently measured by determining the input to the transmission line, and the current at the center of the antenna. Neglecting losses on the transmission line other than ohmic ones, the radiation resistance is the power delivered to the antenna divided by the square of the current.

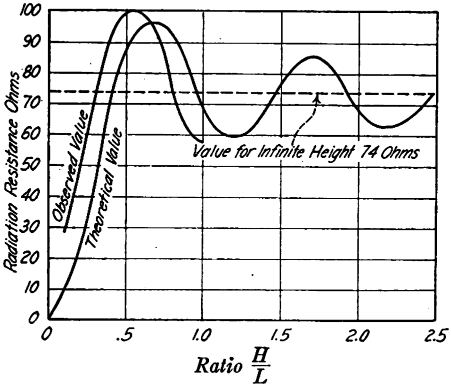

Fig. 22—Variation of radiation resistance with height—horizontal Hertz.

This gives a convenient method for determining the effective resistance of a Hertzian antenna, a method justified by the correlation of the curves in Fig. 22.

When the radiation resistance is known, a measurement can be made of the per cent of power radiated by the feeder. The power delivered to the antenna is obtained by multiplying the square of the current in the center by the radiation resistance. When this is compared with the power delivered to the transmission line by the driver, the curve of Fig. 23 is obtained. This shows conclusively that if the antenna is to be the radiating part of the system, a proper termination is essential.

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[6] Levin and Young, "Radiation Resistance of Vertical Antennas," Proc. I. R. E., **14**, 675–688; October, 1926.

As a final check upon the correctness of the relations deduced, a single-wire transmission line 1200 ft. long was used to feed a half-wave oscillator 50 ft. long. The input power was determined by a calorimeter test. The transmitter was immersed in oil, and a calibration curve of power dissipated vs. rise in temperature was obtained. Knowing the total input to the oscillator and the power lost, the high-frequency output may be determined.

The transmission line was No. 14 copper wire. Such a line has a high I^2R loss and such lengths would not ordinarily be practical. At the frequency used, the ohmic resistance of the 1200-ft. length as computed by the standard formula for skin effect is 48.7 ohms.

The input power used was small but definitely measurable with an accuracy of about 4 per cent. This power was 31.1 watts at a current of 0.320 amperes. The I^2R loss in the feeder was therefore 4.9 watts. The current in the antenna was 0.46 ampere and the antenna resistance was 128 ohms. This high resistance is probably due to losses in the supporting structure which was the University Stadium. This gave the power radiated as 27.2 watts. The power unaccounted for is therefore 1 watt, or about 3 per cent of the total, a result within the precision of measurement. This error is on the side of low rather than high loss in the line. This shows that the power radiated by the feeder, if not zero, is at least very small. With a transmission line of such length, twenty times that of the antenna, radiation, if present, might be expected to be quite prominent. This feeder was tested by J. D. Ryder and E. D. Shipley of the Ohio State University.

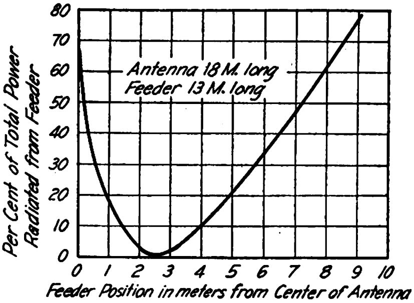

Fig. 23—Per cent of total power radiated from feeder.

It should be noticed that this transmission line is about 12 wavelengths long, so that, as far as radiation problems are concerned, it would correspond at 300 meters to a line 3600 meters in length, or more than two miles.

In running a single-wire line, sharp angles should be avoided, as they cause an irregularity in the distribution of capacity and inductance. On the long feeder, such a sharp bend was at first introduced and the standing waves could not be eliminated until a more gradual bend was made. A similar phenomenon has been noticed on low-frequency power lines when trouble is experienced due to lightning surges piling up at sharp bends. This subject will be studied further.

Since the radiation resistance of the antenna varies markedly with the effective depth of the ground plane, the termination may at times vary with the weather. This would also be true if a two-wire feeder were used.

It is interesting to note that the characteristic impedance of the single-wire line ranges between 600 and 800 ohms for lines of the order of one wavelength long, and therefore is of the same order as a two-wire line. The term "voltage feed" often used to distinguish a single-wire line from a two-wire line or so-called "current feed" is not justified, as the voltage and current ratios of properly terminated lines are about the same on both types. The single-wire line has the advantage of lower I^2R losses. Higher efficiency can be obtained on long lines by loading to increase the impedance, but this is not justifiable on the lengths ordinarily needed.

The results of these and other tests and the mathematical treatment will be set forth more completely in a bulletin by the authors soon to be issued by the Engineering Experiment Station of the Ohio State University.

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CIRCUIT TUNING BY WAVE RESONANCE AND APPLICATIONS TO RADIO RECEPTION*

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Summary—*Theoretical consideration of the wave resonance system of tuning with distributed values of inductance and capacity are given, together with circuit arrangements embodying this method of tuning. It is shown that a high degree of selectivity is obtainable and that it offers an effective method for the elimination of interference. Multiplexing both in the transmission and reception of radio signals can be readily realized.*

IN the present practice of the radio art, circuit tuning is accomplished by a proper combination of localized inductances and capacities; the values of the inductance and capacity being properly chosen to meet the conditions of any particular problem of design. The phenomena connected with this form of tuning have been studied quite exhaustively by many investigators; every conceivable modification of circuit structure has been subjected to careful analysis, and the results made available in many publications.

There is, however, another form of tuning which may be called "Wave Resonance" of which comparatively little (if any) advantage has been taken in the radio art, and which will on investigation offer possibilities in the accomplishment of results which cannot very well be attained by the older method of tuning. The basic principles of this form of tuning were more or less known to theorists, but the application to engineering and particularly to the radio art has been negligible.

It may be well to begin by defining what is meant by "Wave Resonance." It is the resonance condition that obtains when a conductor of distributed inductance and capacity is subjected to electrical oscillations, and either the length of the conductor or the inductance and capacity per unit length of conductor are properly adjusted in relation to the frequency of the oscillations. If an electric impulse is impressed on a conductor of distributed inductance and capacity which is open at the far end, the impulse travels along the conductor and on reaching the open end it is reflected back, returning at the same velocity. If at the time of reaching the starting point an impulse of opposite direction is sent into the conductor, the second impulse adds to the first. If, therefore, alternate impulses are impressed on a conductor and so timed that the intervals between them is the time required for the impulse to travel to and fro on the conductor, a high voltage or current may be produced by small impulses. This, I believe, may be properly called "Wave Resonance," that is, resonance resulting from a wave traveling on a conductor, the length of the conductor being properly adjusted in relation to the frequency or the wavelength of the oscillations.

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This method of tuning offers possibilities for many novel circuit arrangements suitable for various uses in the reception and transmission of radio signals. In the discussion which follows, circuit systems embodying this method of tuning are analyzed. It is shown that a high degree of selectivity is obtainable, and that it offers an effective method for the elimination of interference. Also multiplexing, both in the transmission and reception of radio signals, can be readily realized by a circuit system utilizing this method of tuning.

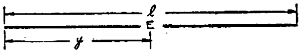

Fig. 1

ANTENNA TUNING

We shall consider first the case of a simple free antenna open at both ends, and examine the conditions for tuning under the action of a signal e.m.f. impressed uniformly at every point of the antenna. For a conductor of distributed inductance and capacity open at both ends and an e.m.f. E impressed at some intermediate point distance y from the left end of the conductor, the voltages V_1 and V_2 at points on the conductor to the right side of E and left side of E are given by the following equations[1]

$$V_1 = \frac{E \sinh ky}{\sinh kl} \cosh k(l-x), \qquad V_2 = \frac{-E \sinh k(l-y)}{\sinh kl} \cosh kx. \tag{1}$$

The total voltage, that due to the e.m.f. acting at every point on the antenna, is readily obtainable by integrating the above expressions with respect to y. Since for every point voltage we get voltage waves on either side of the point of application, the total voltage is the integrated value of the sum of the voltages given by (1). That is

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[1] Louis Cohen. "Heaviside's Electrical Circuit Theory," p. 76.

$$V_t = E\int_0^l \frac{\sinh ky \cosh k(l-x) - \sinh k(l-y)\cosh kx}{\sinh kl}\,dy$$

$$= \frac{\dfrac{El}{Kl}\Big\{(\cosh kl - 1)\left[\cosh k(l-x) - \cosh kx\right]\Big\}}{\sinh kl}. \tag{2}$$

For $x=0$,

$$\left.\begin{aligned} &V_t = \frac{\dfrac{El}{kl}(\cosh kl - 1)^2}{\sinh kl}. \\ \text{For } x=l, \qquad &V_t = \frac{-\dfrac{El}{kl}(\cosh kl - 1)^2}{\sinh kl}. \end{aligned}\right\} \tag{3}$$

At $x=l/2$, $V_t=0$, a voltage node which we would expect.

In the above expressions

$$\left.\begin{aligned} k &= \sqrt{(L\omega+R)cj\omega} = \alpha + j\beta, \\ \alpha &= \sqrt{\tfrac{1}{2}c\omega\left\{\sqrt{L^2\omega^2+R^2} - L\omega\right\}}, \\ \beta &= \sqrt{\tfrac{1}{2}c\omega\left\{\sqrt{L^2\omega^2+R^2} + L\omega\right\}} \end{aligned}\right\} \tag{4}$$

For high frequencies $L^2\omega^2$ is large in comparison with R^2, and for this case to a high degree of approximation

$$\left.\begin{aligned} \alpha &= \tfrac{1}{2}R\sqrt{\frac{C}{L}}, \\ \beta &= \omega\sqrt{LC}. \end{aligned}\right\} \tag{5}$$

For $kl = \alpha l + j\beta l$, we have

$$\left.\begin{aligned} \sinh kl &= \tfrac{1}{2}(e^{kl} - e^{-kl}) = \tfrac{1}{2}\left\{\cos\beta l(e^{\alpha l} - e^{-\alpha l}) + j\sin\beta l(e^{\alpha l} + e^{-\alpha l})\right\}, \\ \cosh kl &= \tfrac{1}{2}(e^{kl} + e^{-kl}) = \tfrac{1}{2}\left\{\cos\beta l(e^{\alpha l} + e^{-\alpha l}) + j\sin\beta l(e^{\alpha l} - e^{-\alpha l})\right\}. \end{aligned}\right\} \tag{6}$$

For small values of αl which would be the case for a short conductor as that of an antenna, the above reduce to

$$\left.\begin{aligned} \sinh kl &= \alpha l \cos\beta l + j\sin\beta l, \\ \cosh kl &= \cos\beta l + j\alpha l \sin\beta l. \end{aligned}\right\} \tag{7}$$

Introducing these values in either of equations (3) we get the following expressions for the voltage at either end of the conductor

$$V_t = \frac{\dfrac{El}{j\beta l}(\cos \beta l + j\alpha l \sin \beta l - 1)^2}{\alpha l \cos \beta l + j \sin \beta l}. \tag{8}$$

A voltage resonance condition obtains for this case when $\beta l = n\pi$, $n = 1, 3, 5, \cdots$, that is $\cos \beta l = -1$, and $\sin \beta l = 0$. This gives:

$$V_{\text{res.}} = \frac{4El}{jn\pi\alpha l} = \frac{8El\sqrt{\dfrac{L}{C}}}{jn\pi Rl}. \tag{9}$$

If Rl is of the order of magnitude of 3 ohms and $\sqrt{\dfrac{C}{L}} = 300$, reasonable values, we get for the resonance voltage $V_{res} = \dfrac{800El}{\pi}$.

Formula (9) for the resonance voltage on a line is of the same form as that of the voltage across the condenser of an oscillating circuit, and the voltage rise is also of the same order of magnitude that obtains in a closed oscillatory circuit. The sharpness of tuning, however, is more marked than would be the case for a closed circuit. This can be readily seen by a simple example. Suppose the frequency is slightly different from the resonance frequency; that is $\beta l = \pi + x$ where x is a small quantity, obviously the value of $\cos \beta l$ will be changed but little and $\sin \beta l$ will change to $-\sin x = -x$. The denominator in (8) will be changed from $-\alpha l$ to $-\alpha l - jx$. If $x = 2$ deg. representing a change in frequency of about one per cent, the denominator in (8) will be changed from $-\alpha l = \dfrac{1}{200}$, at resonance, to $-\dfrac{1}{200} - j\dfrac{4\pi}{360} = -0.005$ $-j\,0.035$. The change indicated is from 0.005 to 0.035. The denominator is increased seven fold giving a voltage reduction to one seventh of the resonance value for a one per cent change in frequency; a much greater change than would be the case for a closed circuit.

For the condition $\beta l = n\pi$, we have on substituting $\omega\sqrt{LC}$ for β,

$$f = \frac{n}{2l\sqrt{LC}}. \tag{10}$$

In all cases of wave motion the wave velocity V is connected with the wavelength λ and the frequency f in the relation given by the formula

$$V = \lambda f,$$

but

$$V = \frac{1}{\sqrt{LC}},$$

therefore

$$f = \frac{1}{\lambda\sqrt{LC}}.$$

Substituting this in (10), we get

$$\lambda = \frac{2l}{n},$$

or

$$l = \frac{n\lambda}{2}.$$

For a conductor open at both ends, the resonance condition obtains when the length of the conductor is an odd multiple of half wavelengths.

It can be readily shown that for a conductor grounded at one end the resonance condition obtains when the length of the conductor is adjusted for an odd multiple of quarter wavelengths of the waves developed on it.

It is clear from what preceded that effective selective tuning may be accomplished by the use of open oscillators. To use straight conductors, however, would be impractical on account of the great lengths that would be required, and also because of the difficulties in the matter of adjustment and of the means for transferring the energy to some suitable receiving instruments. In what follows, methods are discussed by which this principle may be utilized in a practical system for the reception and transmission of radio signals. It is also shown that multiplexing, transmitting, and receiving can be accomplished by this method of tuning.

WAVE CONDUCTORS

To make use of wave resonance tuning in a practical way, a wave conductor must be made available that will meet the conditions of practical dimensions, and embodying convenient means of adjustment for different frequencies. This is realized in the design of a wave conductor which consists of a solenoidal coil mounted on a metal plate. A photograph of a typical wave conductor used in the experiments is shown in Fig. 2. The dimensions of the coils used for the broadcast frequency range are 8 in. long and 2 in. in diameter, wound with a suitable number of turns of wire giving the required length of conductor, and at the same time keeping the dimensions within practical limits. The dimensions of the coil as well as the number of turns may be varied to meet particular requirements of design in the matter of frequency range. By varying the distance separation of the movable plate from the coil, the distributed capacity and, to some extent, the distributed inductance are varied, affording a convenient means for adjustment. In some circuit systems the metal plate is grounded and in others ungrounded, depending upon the particular circuit arrangement in which it is used.

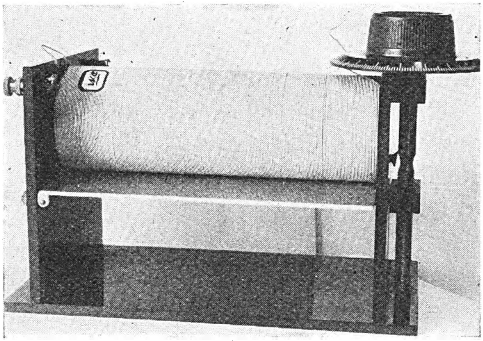

Fig. 2

The assumption that a conductor of this kind possesses the characteristics of uniform distributed inductance and capacity may not be strictly true, but sufficiently accurate for the formulation of a working theory. A more comprehensive theory which is to take cognizance of the variation of the inductance and capacity along the length of the conductor would introduce mathematical difficulties and lead to complicated formulas which may obscure the physical aspects of the problem. As a first formulation of the theory, therefore, it is considered advisable to make use of the assumption of uniform inductance and capacity which simplifies the discussion of the problem.

ANTENNA AND WAVE CONDUCTOR IN SERIES

In discussions of antenna receiving circuits it is generally assumed that the antenna may be considered as equivalent to a localized capacity in series with an emf which is that produced by the received signals acting on the entire length of the antenna. This assumption is permissible when the antenna is short in comparison with the signal wavelength. For high frequencies, where the antenna length may be comparable to the wavelength of the signals, a serious error may be introduced in considering the antenna the equivalent of a localized capacity. It is preferable, therefore, to consider the problem in the more general form taking into accunt the distributed antenna characteristic, and where permissible the expressions arrived at will reduce to the simpler forms which would be obtained by considering the antenna as a localized capacity.

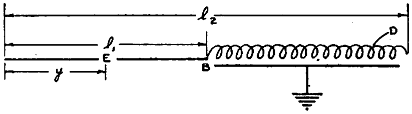

Fig. 3

We shall consider first the voltage and current distribution in a wave conductor which is connected in series with an antenna as shown in Fig. 3. Designate by L_1, R_1, C_1, and L_2, R_2, C_2 the inductance, resistance, and capacity per unit length of antenna and wave conductor respectively, and let

$$\left.\begin{aligned} k_1&=\sqrt{(L_1j\omega+R_1)C_1j\omega_1}=\alpha_1+j\beta_1,\\ k_2&=(L_2j\omega+R_2)C_2j\omega=\alpha_2+j\beta_2.\end{aligned}\right\}\qquad(12)$$

For a point emf E acting at an intermediate point of the antenna distance y from the open end of the antenna we have the following expressions for the currents and voltages.

For the antenna to the left side of E

$$\left.\begin{aligned} I_1&=A_1e^{k_1x}+B_1e^{-k_1x},\\ V_1&=\frac{k_1}{C_1j\omega}\{A_1e^{k_1x}-B_1e^{-k_1x}\}.\end{aligned}\right\}\qquad(13)$$

For the antenna to the right side of E

$$\left.\begin{aligned} I_2&=A_2e^{k_1(x-y)}+B_2e^{-k_1(x-y)},\\ V_2&=\frac{k_1}{C_1j\omega}\{A_2e^{k_1(x-y)}-B_2e^{-k_1(x-y)}\}.\end{aligned}\right\}\qquad(14)$$

For the wave conductor

$$\left.\begin{aligned} I_3&=A_3e^{k_2(x-l_1)}+B_3e^{-k_2(x-l_1)}.\\ V_3&=-\frac{k_2}{C_2j\omega}\{A_3e^{k_2(x-l_1)}-B_3e^{-k_2(x-l_1)}\}.\end{aligned}\right\}\qquad(15)$$

The terminal conditions from which the six constants of the above three sets of equations are determined are as follows

$$\left.\begin{array}{ll} x=0; & I_1=0. \\ x=l_2; & I_3=0. \\ x=l_1. & I_2=I_3 \text{ and } V_2=V_3. \\ x=y; & I_1=-I_2 \text{ and } V_1-V_2=E. \end{array}\right\} \qquad (16)$$

These give the following six equations

$$\left.\begin{array}{r} A_1+B_1=0, \\ A_3e^{k_2(l_2-l_1)}+B_3e^{-k_2(l_2-l_1)}=0, \\ A_2e^{k_1(l_1-y)}+B_2e^{-k_1(l_1-y)}=A_3+B_3, \\ \dfrac{k_1}{C_1j\omega}\left\{A_2e^{k_1(l_1-y)}-B_2e^{-k_1(l_1-y)}\right\}=\dfrac{k_2}{C_2j\omega}(A_3-B_3), \\ A_1e^{k_1y}+B_1e^{-k_1y}=-A_2-B_2, \\ \dfrac{k_1}{C_1j\omega}\left\{A_1e^{k_1y}-B_1e^{-k_1y}+A_2-B_2\right\}=E. \end{array}\right\} \qquad (17)$$

From these equations all the constants are readily determined, which on substitution in (13), (14), and (15) give the complete solutions for the current and voltage distribution in the antenna and wave conductor. The main interest in this study, however, is the current and voltage distribution in the wave conductor, and we write therefore only the expressions for A_3 and B_3 which are as follows

$$A_3=\frac{E\left\{e^{k_1y}-e^{-k_1y}\right\}e^{-k_2(l_2-l_1)}}{\dfrac{k_1}{C_1j\omega}(e^{k_1l_1}+e^{-k_1l_1})(e^{k_2(l_2-l_1)}-e^{-k_2(l_2-l_1)})+\dfrac{k_2}{C_2j\omega}(e^{k_1l_1}-e^{-k_1l_1})(e^{k_2(l_2-l_1)}+e^{-k_2(l_2-l_1)})} \qquad (18)$$

$$B_3=\frac{-E\left\{e^{k_1y}-e^{-k_1y}\right\}e^{k_2(l_2-l_1)}}{\dfrac{k_1}{C_1j\omega}(e^{k_1l_1}+e^{-k_1l_1})(e^{k_2(l_2-l_1)}-e^{-k_2(l_2-l_1)})+\dfrac{k_2}{C_2j\omega}(e^{k_1l_1}-e^{-k_1l_1})(e^{k_2(l_2-l_1)}+e^{-k_2(l_2-l_1)})}$$

$$\left.\begin{array}{l} I_3=\dfrac{E(e^{k_1y}-e^{-k_1y})(e^{k_2(x-l_2)}-e^{-k_2(x-l_2)})}{\dfrac{k_1}{C_1j\omega}(e^{k_1l_1}+e^{-k_1l_1})(e^{k_2(l_2-l_1)}-e^{-k_2(l_2-l_1)})+\dfrac{k_2}{C_2j\omega}(e^{k_1l_1}-e^{-k_1l_1})(e^{k_2(l_2-l_1)}+e^{-k_2(l_2-l_1)})}, \\ V_3=\dfrac{-\dfrac{k_2}{C_2j\omega}E(e^{k_1y}-e^{-k_1y})(e^{k_2(x-l_1)}+e^{-k_2(x-l_1)})}{\dfrac{k_1}{C_1j\omega}(e^{k_1l_1}+e^{-k_1l_1})(e^{k_2(l_2-l_1)}-e^{-k_2(l_2-l_1)})+\dfrac{k_2}{C_2j\omega}(e^{k_1l_1}-e^{-k_1l_1})(e^{k_2(l_2-l_1)}+e^{-k_2(l_2-l_1)})} \end{array}\right. \qquad (19)$$

Substituting these values in (15), we have the complete expressions for the current and voltage at any point on the wave conductor due to an emf acting at one point on the antenna.

For the total signal e.m.f. acting at every point on the antenna, we have

$$I_3' = \int_0^{l_1} I_3 dy,$$

$$V_3' = \int^{l_1} V_3 dy.$$

Introducting the values of I_3 and V_3 from (19) and integrating, we get

$$\left.\begin{aligned} I_3' &= \frac{\dfrac{El_1}{k_1 l_1}(e^{k_1 l_1} + e^{-k_1 l_1} - 2)(e^{k_2(l_2-x)} - e^{-k_2(l_2-x)})}{\dfrac{k_1}{C_1 j\omega}(e^{k_1 l_1} + e^{-k_1 l_1})(e^{k_2(l_2-l_1)} - e^{-k_2(l_2-l_1)}) + \dfrac{k_2}{C_2 j\omega}(e^{k_1 l_1} - e^{-k_1 l_1})(e^{k_2(l_2-l_1)} + e^{-k_2(l_2-l_1)})}, \\ V_3' &= \frac{\dfrac{k_2}{C_2 j\omega}\dfrac{El_1}{k_1 l_1}(e^{k_1 l_1} + e^{-k_1 l_1} - 2)(e^{k_2(l_2-x)} + e^{-k_2(l_2-x)})}{\dfrac{k_1}{C_1 j\omega}(e^{k_1 l_1} + e^{-k_1 l_1})(e^{k_2(l_2-l_1)} - e^{-k_2(l_2-l_1)}) + \dfrac{k_2}{C_2 j\omega}(e^{k_1 l_1} - e^{-k_1 l_1})(e^{k_2(l_2-l_1)} + e^{-k_2(l_2-l_1)})} \end{aligned}\right\} \quad (20)$$

These are the most general expressions applicable for any antenna and any signal frequency. For the case, however, where the length of the antenna is only a small fraction of the signal wavelength, approximations may be introduced which simplify considerably the formulas. The antenna attenuation may be always neglected, and in that case we may put

$$e^{k_1 l_1} + e^{-k_1 l_1} = 2 \cos \beta_1 l_1,$$

$$e^{k_1 l_1} - e^{-k_1 l_1} = 2j \sin \beta_1 l_1.$$

Now $\beta_1 l_1 = 2\pi f \sqrt{L_1 C_1}\, l_1 = 2\pi l_1/\lambda$, and for small values of l/λ, we have approximately

$$\cos \beta_1 l_1 = 1 - \frac{\beta_1 l_1{}^2}{2},$$

$$\sin \beta_1 l_1 = \beta_1 l_1.$$

Using these approximations, equations (20) reduce to

$$\left.\begin{aligned} I_3' &= \frac{\dfrac{E_0}{2}(e^{k_2(l_2-x)} - e^{-k_2(l_2-x)})}{\dfrac{1}{j\omega C_0}(e^{k_2(l_2-l_1)} - e^{-k_2(l_2-l_1)}) + \dfrac{k_2}{C_2 j\omega}(e^{k_2(l_2-l_1)} + e^{-k_2(l_2-l_1)})}, \\ V_3' &= \frac{\dfrac{E_0}{2}\dfrac{k_2}{C_2 j\omega}(e^{k_2(l_2-x)} + e^{-k_2(l_2-x)})}{\dfrac{1}{j\omega C_0}(e^{k_2(l_2-l_1)} - e^{-k_2(l_2-l_1)}) + \dfrac{k_2}{C_2 j\omega}(e^{k_2(l_2-l_1)} + e^{-k_2(l_2-l_1)})}. \end{aligned}\right\} \quad \textbf{(21)}$$

Where $E_0=El_1$, the total emf acting on the antenna, and $C_0=Cl_1$, the total capacity of the antenna.

These are precisely the expressions that would result by considering the antenna as a localized capacity to begin with, but it must be remembered that they are only valid for the case l_1/λ small. For short waves the complete formulas (20) must be used.

If the wave conductor is of short length, which would be generally the case, $\alpha_2(l_2-l_1)$ is very small, and to a high degree of approximation,

$$\begin{aligned} e^{k_2(l_2-l_1)}+e^{-k_2(l_2-l_1)} &= 2\cos\beta_2(l_2-l_1)+2j\alpha_2(l_2-l_1)\sin\beta_2(l_2-l_1), \\ e^{k_2(l_2-l_1)}-e^{-k_2(l_2-l_1)} &= 2\alpha_2(l_2-l_1)\cos\beta_2(l_2-l_1)+2j\sin\beta_2(l_2-l_1). \end{aligned} \tag{22}$$

Introducing these approximations in (21), the expression for the voltage on the wave conductor reduces to

$$V_3'=\frac{\dfrac{E_0}{2}\sqrt{\dfrac{L_2}{C_2}}\{\cos\beta_2(l_2-x)+j\alpha_2(l_2-x)\sin\beta_2(l_2-x)\}}{\dfrac{1}{j\omega C_0}\{\alpha_2(l_2-l_1)\cos\beta_2(l_2-l_1)+j\sin\beta_2(l_2-l_1)\}+\sqrt{\dfrac{L_2}{C_2}}\{\cos\beta_2(l_2-l_1)+j\alpha_2(l_2-l_1)\sin\beta_2(l_2-l_1)\}} \tag{23}$$

For $\beta_2(l_2-l_1)=\pi/2+z$, the above reduces to

$$V_3'=\frac{\dfrac{E_0}{2}\sqrt{\dfrac{L_2}{C_2}}\{\cos\beta_2(l_2-x)+j\alpha_2(l_2-x)\sin\beta_2(l_2-x)\}}{\dfrac{1}{j\omega C_0}\{-\alpha_2(l_2-l_1)\sin z+j\cos z\}+\sqrt{\dfrac{L_2}{C_2}}\{-\sin z+j\alpha_2(l_2-l_1)\cos z\}}. \tag{24}$$

Assume now that an adjustment is made in $\beta_2(l_2-l_1)$, which is accomplished in practice by changing the distance between the metal plate and the coil of the wave conductor, so that

$$\frac{1}{\omega C_0}\cos z=\sqrt{\frac{L_2}{C_2}}\sin z, \tag{25}$$

then (24) reduces to

$$V_3'=\frac{\dfrac{E_0}{2}\sqrt{\dfrac{L_2}{C_2}}\{\cos\beta_2(l_2-x)+j\alpha_2(l_2-x)\sin\beta_2(l_2-x)\}}{j\alpha_2(l_2-l_1)\left\{\dfrac{1}{\omega C_0}\sin z+\sqrt{\dfrac{L_2}{C_2}}\cos z\right\}}. \tag{26}$$

which is the resonance condition. At $x=l_2$, the end of the wave conductor,

$$V_3=\frac{\frac{E_0}{2}\sqrt{\frac{L_2}{C_2}}}{j\alpha_2(l_2-l_1)\left\{\frac{1}{C_0\omega}\sin z+\sqrt{\frac{L_2}{C_2}}\cos z\right\}}. \tag{27}$$

In practice $\sqrt{L_2/C_2}$ is large compared with $1/C_0\omega$. We may assume as reasonable values $\sqrt{L_2/C_2}=10^4$, and $1/C_0\omega=200$, and for these values we have by (25),

$$\tan z=0.02,$$
$$z=1^\circ\,9'.$$

Sin z is very small and cos z is practically unity, and in that case

$$V_3=\frac{E_0\sqrt{\frac{L_2}{C_2}}}{R_2(l_2-l_1)}. \tag{28}$$

At a point on the wave conductor distance x determined by the relation $\beta_2(l_2-x)=\frac{\pi}{2}$, the voltage is given by, in accordance with (26)

$$V_3=\frac{\frac{E_0}{2}\sqrt{\frac{L_2}{C_2}}(l_2-x)}{(l_2-l_1)\left\{\frac{1}{C_0\omega}\sin z+\sqrt{\frac{L_2}{C_2}}\cos z\right\}},$$

$$=\frac{E_0}{2}\,\frac{l_2-x}{l_2-l_1}\text{ approximately}. \tag{29}$$

The voltage at this point is very small compared with the voltage at the end of the wave conductor. If we use the value $\sqrt{L_2/C_2}=10^4$, and assume the total resistance of the wave conductor to be 40 ohms, the ratio of the voltages, that at the end of the wave conductor and at

$$x=l_2-\frac{\pi}{2\beta_2}=l_2-\frac{\lambda}{4},$$

$$\frac{V_{3l_2}}{V_3}=\frac{2\times10^4}{40}=500.$$

At the point $x=l_2-\lambda/4$, the voltage wave has a nodal point. If it were not for the resistance of the wave conductor, the voltage at that point would be of zero value. The voltage wave form is shown graphically in Fig. 4.

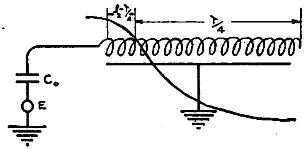

Fig. 4

It is quite evident that a double tuning is accomplished in the process of adjusting the wave conductor; the small part of the wave conductor, $l_2-\lambda/4$, tunes out the antenna capacity reactance, and on the balance of the wave conductor a quarter wavelength is established. It must be remembered that all of this is accomplished by one adjustment, it is merely a matter of varying β_2, which is effected by changing the distance separation of the metal plate and the wave conductor until the condition $\beta l=\pi/2+z$ is realized, which is the resonance condition.

That a high degree of selectivity is obtainable by this system of tuning will appear from the consideration of a numerical example. For the values chosen, $\sqrt{L_2/C_2}=10^4$ and $1/C_0\omega=200$, the resonance condition obtains when $\beta_2(l_2-l_1)=91$ deg. 9 min., and for that value the resonance voltage at the end of the wave conductor is by (28), assuming $R_2(l_2-l_1)=40$ ohms.

$$V_{3,\,\text{res.}}=\frac{E\times 10^4}{40}=250E.$$

For $\beta_2(l_2-l_1)=90$ deg., about one per cent reduction in frequency.

$$\cos\beta_2(l_2-l_1)=0; \qquad \sin\beta_2(l_2-l_1)=1,$$

and for these values

$$V_3=\frac{E}{2}\times\frac{10^2}{200+j20}=25E.$$

For $\beta_2(l_2-l_1)=92$ deg., about one per cent increase in frequency, we have

$$\cos\beta_2(l_2-l_1)=-0.035,$$

$$\sin\beta_2(l_2-l_1)=0.9994,$$

For these values,

$$V_3 = \frac{\frac{E_0}{2} \times 10^4}{-j200\{-20 \times 0.035 \times 10^{-4} + j0.9994\} + 10^4\{-0.035 + j40 \times 10^{-4}\}}$$

$$= \frac{E_0}{2} \times \frac{10^4}{(200-350)} = 33E.$$

A change in frequency of one per cent on either side of the resonance frequency produces a reduction in voltage to about ten per cent of the resonance voltage, which shows clearly the tuning effectiveness of the system.

This form of tuning may be utilized in various ways in circuit systems for the reception and transmission of radio signals.

Special consideration, however, is required in using this system for very short waves. In that case the more complete formula (20) should be used, which gives for the voltage at the end of the wave conductor, $x = l_2$,

$$V_3 = \frac{\frac{El_1}{k_1 l_1} \frac{k_2}{C_2 j\omega} \{e^{k_1 l_1} + e^{-k_1 l_1} - 2\}}{\frac{k_1}{C_1 j\omega}(e^{k_1 l_1} + e^{-k_1 l_1})(e^{k_2(l_2-l_1)} - e^{-k_2(l_2-l_1)}) + \frac{k_2}{C_2 j\omega}(e^{k_1 l_1} - e^{-k_1 l_1})(e^{k_2(l_2-l_1)} + e^{-k_2(l_2-l_1)})} \tag{30}$$

Using the approximations given by (22) the above transforms to

V_3 = a fraction, the numerator of which is $\frac{1}{2} \frac{El_1}{k_1 l_1} \frac{k_2}{C_2 j\omega}\{\cos \beta_1 l_1$

$+ j\alpha_1 l_1 \sin \beta_1 l_1 - 1\}$ and the denominator of which is $\frac{k_1}{C_1 j\omega}(\cos \beta_1 l_1$ (31)

$+ j\alpha_1 l_1 \sin \beta_1 l_1)\{\alpha_2(l_2 - l_1) \cos \beta_2(l_2 - l_1) + j \sin \beta_2(l_2 - l_1)\}$

$+ \frac{k_2}{C_2 j\omega}(\alpha_1 l_1 \cos \beta_1 l_1 + j \sin \beta_1 l_1)\{\cos \beta_2(l_2 - l_1) + j\alpha_2(l_2 - l_1) \sin \beta_2(l_2 - l_1)\}$.

Tuning is accomplished by adjusting the wave conductor to satisfy the condition

$$\frac{k_1}{C_1 j\omega} \cos \beta_1 l_1 \sin \beta_2(l_2 - l_1) = -\frac{k_2}{C_2 j\omega} \sin \beta_1 l_1 \cos \beta_2(l_2 - l_1),$$

or

$$\tan \beta_1 l_1 = -\sqrt{\frac{L_1 C_2}{L_2 C_1}} \tan \beta_2(l_2 l_1). \tag{32}$$

For this condition, neglecting small quantities of the second order, that is terms in $\alpha^2 l^2$, equation (31) reduces to

$$V_3 = \frac{\frac{1}{2}\frac{El_1}{k_1 l_1}\frac{k_2}{C_{2j}\omega}\left\{\cos\beta_1 l_1 + j\alpha_1 l_1 \sin\beta_1 l_1 - 1\right\}}{\cos\beta_1 l_1 \cos\beta_2(l_2 - l_1)\left\{\frac{k_1}{C_{1j}\omega}\alpha_2(l_2 - l_1) + \frac{k_2}{C_{2j}\omega}\alpha_1 l_1\right\} - \sin\beta_1 l_1 \sin\beta_2(l_2 - l_1)\left\{\frac{k_1}{C_{1j}\omega}\alpha_1 l_1 + \frac{k_2}{C_{2j}\omega}\alpha_2(l_2 - l_1)\right\}} \tag{33}$$

Each of the terms in the denominator contains an αl factor which is a very small quantity, and hence the resonance voltage rise at the end of the wave conductor is great.

It is to be observed that in accordance with (32), the tuning of the wave conductor for the resonance condition is not only a matter of frequency but also of antenna length. For the same frequency the wave conductor will tune differently for antennas of different lengths.

Another point to be noticed in connection with this system for short wave tuning is the special case when the length of the antenna is a whole wavelength of the signals, which makes $\beta_1 l_1 = 2\pi$, and in that case the numerator of (33) is practically zero, and there is no voltage transfer at all to the wave conductor. That this should be so is evident from a consideration of equations (3) which show that for a distributed emf over a conductor a wavelength long the voltages are zero at the ends of the conductor, and it would follow that by connecting a wave conductor to the end of an antenna when this condition exists (zero voltage at the end), there would be no energy transfer to the wave conductor. It is obvious from this that in using this system of tuning care must be taken to avoid the condition of having the length of the antenna approximate a whole wavelength or a multiple of wavelengths of the signals.

RECEPTION OF SIGNALS

The possibility of the adoption of wave resonance tuning for the reception of radio signals is quite obvious from the preceding discussion, provided suitable means are introduced for transferring to a receiving device the voltage or current energy from the wave conductor. This can be accomplished in various ways; it is largely a matter of coupling. The simplest way is to couple a closed circuit, preferably tuned, to the wave conductor, and the voltage developed across the condenser of the closed circuit applied to the grid filament of a vacuum tube in the usual way. If desired, additional tuning may be had by the use of a tuned radio-frequency amplifier, and in that case the wave conductor tuning serves to superimpose another degree of tuning, giving a remarkably high degree of selectivity and this without any loss in efficiency. In fact it may be used in connection with any of the well known types of commercial receivers. In practice it is found that a sufficient degree of coupling is obtained by placing the wave conductor close to the receiving set, the coupling being largely electrostatic. A better arrangement, however, is to connect the metal plate of the wave conductor to the receiving set as shown in Fig. 5.

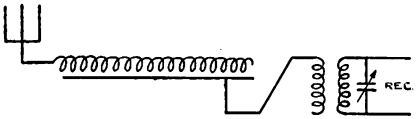

Fig. 5

In this case the metal plate serves a double purpose; that of adjusting the wave conductor, and as a coupling means. I had the use of a Stromberg-Carlson set in which the primary is untuned, the metal plate of the wave conductor connected to one terminal of the primary coil, and the other terminal open. The coupling from the primary to the tuned secondary circuit is obviously electrostatic. Very good results in the matter of volume and selectivity were obtained with this arrangement.

A somewhat startling phenomenon was observed in connection with reception tests by the arrangement of Fig. 5, that, even when grounding the plate of the wave conductor it was possible to receive signals from distant stations though with loss in volume. It would seem that since grounding the metal plate grounds the receiver, no signal reception at all could be expected. The fact is, however, that signals were received with the grounded plate arrangement, and it was also noted that the tuning of the wave conductor was sharpened to a remarkable degree, requiring an extremely delicate adjustment. The only explanation that can be offered is that the ground is not a zero potential surface, as it is customary to assume but perhaps more satisfactory explanations will be forthcoming in the future.

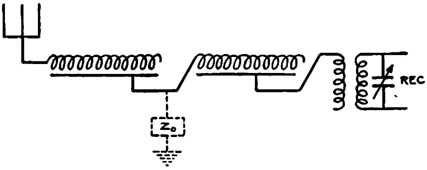

Fig. 6

A modification of Fig. 5 by which the selectivity is considerably enhanced is shown in Fig. 6, in which two wave conductors connected in tandem are used, each separately tuned. It is clear that by this arrangement a progressive tuning is effected. An additional improvement is obtained by grounding either directly, or through an impedance the junction point of the two wave conductors as indicated by the dotted lines in the figure.

ELIMINATION OF INTERFERENCES

Tuning by wave resonance is well adapted for the elimination of interferences. It was shown in the preceding discussion that a wave conductor connected in series with an antenna can be made to respond resonantly by proper adjustment to signals of a particular frequency, and it should follow from this that it should also function efficiently as an energy absorber of an interfering signal of that particular frequency. That this is actually so was demonstrated experimentally, and accomplished in the following manner. The receiver is connected to the antenna in the usual manner, and a wave conductor connected to the junction point of antenna and receiver as shown in Fig. 7. By the proper adjustment of the wave conductor to the frequency of any particular interfering signal, a voltage node is established for that frequency at the junction point of antenna and receiver, and consequently very little if any signal current of that particular frequency will pass into the receiver. The following brief mathematical discussion will make this evident.

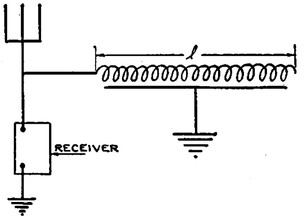

Fig. 7

Let Z_1 denote the antenna reactance, and Z_2 the receiver reactance; I_1 and I_2 the currents in the antenna and receiver, respectively. The expressions for the current and voltage of wave conductors are as follows:

$$\left.\begin{aligned} I &= Ae^{kx}+Be^{-kx}, \\ V &= -\frac{k}{Cj\omega}\left\{Ae^{kx}-Be^{-kx}\right\}. \end{aligned}\right\} \qquad (34)$$

The constants A and B are determined from the following terminal conditions:

For $x=l;$ $I=o,$

For $x=0;$ $V_0=E-Z_1I_1.$

also,

$$I_1=I_2+I_{x=0},$$

and

$$I_2=\frac{V_{x=0}}{Z_2}.$$

These conditions give,

$$\left.\begin{aligned} &Ae^{kl}+Be^{-kl}=0, \\ &-\frac{k}{Cj\omega}(A-B)=E+\frac{Z_1}{Z_2}\frac{k_1}{Cj\omega}(A-B)-Z_1(A+B). \end{aligned}\right\} \quad (35)$$

from which the values of A and B are readily determined. Introducing these in (34), we get the complete expression for the voltage on the wave conductor as follows:

$$V=\frac{-E\frac{k}{Cj\omega}\left\{e^{k(l-x)}+e^{-k(l-x)}\right\}}{Z_1(e^{kl}-e^{-kl})+\frac{k}{Cj\omega}\left(1+\frac{Z_1}{Z_2}\right)(e^{kl}+e^{-kl})}. \quad (36)$$

At $x=0,$

$$V_{x=0}=\frac{-E\frac{k}{Cj\omega}(e^{kl}+e^{-kl})}{Z_1(e^{kl}-e^{-kl})+\frac{k}{Cj\omega}\left(1+\frac{Z_1}{Z_2}\right)(e^{kl}+e^{-kl})}. \quad (37)$$

Using the approximations given by (22), the above reduces to

$$V_{x=0}=\frac{-E\frac{k}{Cj\omega}\{\cos\beta l+j\alpha l\sin\beta l\}}{Z_1(\alpha l\cos\beta l+j\sin\beta l)+\frac{k}{Cj\omega}\left(1+\frac{Z_1}{Z_2}\right)(\cos\beta l+j\alpha l\sin\beta l)}. \quad (38)$$

When the wave conductor is adjusted for the condition $\beta l=\pi/2$, the bracket term in the numerator of (38) reduces to αl, which is a very small quantity of the order of magnitude of 0.001 or 0.002 depending on the resistance and the ratio of C/L of the wave conductor, and therefore the interfering signal of the particular frequency for which the wave conductor is in adjustment to the condition $\beta l=\pi/2$ can have very little effect on the receiver. For signals of other frequencies, for example differing five per cent in frequency, $\beta l = 1.05\pi/2 = 94.5$ deg., and cos $\beta l = -0.078$, the voltage would be any where from 39 to 78 times greater than the voltage of the interfering signal for the same emf acting on the antenna. It is obvious that by connecting several wave conductors, each separately adjusted for the frequency of a different interfering signal, any number of interfering signals may be eliminated simultaneously, and this without any appreciable reduction in the efficiency of reception of signals of other frequencies. This is amply borne out by experimental tests in connection with various types of commercial receivers.

This immediately suggests a method for further increasing the selectivity of reception by the wave resonance tuning method. It consists in replacing the receiver of Fig. 7 by a leak to ground, and coupling the receiver to the wave conductor in the manner indicated in the preceding section. It is necessary, however, to introduce an inductance in the antenna to tune the antenna separately. For the resonance condition then we shall have

$$V = \frac{-E\dfrac{k}{Cj\omega}}{R_1 j + \dfrac{k}{Cj\omega}\left(1+\dfrac{R_1}{Z_2}\right)j\alpha l}\,.$$

If Z_2 is a resistance of a value R_1, the above reduces to

$$V = \frac{-E\sqrt{\dfrac{L}{C}}}{j(R_1+Rl)},$$

where Rl is the total resistance of wave conductor. For $\sqrt{L/C} = 10^4$, $R_1 = 10$ and $Rl = 40$,

$$V = 200E.$$

For of resonance, say 5 per cent $\beta l = 85.5$ deg., cos $\beta l = 0.0785$, sin $\beta l = 0.997$, and Z_1 may change from 10 to 100. For this case

$$V = \frac{-E\times 10^4}{j100(j0.997)+10^4(1+j10)(0.0785)},$$

$$= -\frac{E\times 10^4}{10^4\times 0.785} = 1.3E.$$

A reduction in voltage to 1.3/200 of the resonance value, which would indicate a very high degree of selectivity. It is obvious that this method may be extended to the use of a three-quarter wavelength conductor establishing another voltage nodal point at the half wavelength distance, and inserting at that point another leak to ground. This would still further increase the selectivity of the system.

MULTIPLEX RECEPTION

The possibility of utilizing wave resonance tuning for multiplex reception naturally suggests itself. If, as has been indicated, it is possible to eliminate several interfering signals of different frequencies by the use of several wave conductors each adjusted to a frequency of one of the interfering signals, then it should be possible to reverse the process by coupling receivers to the wave conductors through each of which a signal of one frequency only (for which it is in adjustment) is transmitted to the corresponding receiver. This has been accomplished in a highly satisfactory manner. An outline of the theory is given in what follows.

It was already shown that for short antennas, that is, short in comparison with the wavelength of the signals, it is fairly accurate to consider the antenna as a localized capacity, and we shall make use of this approximation in the following discussion which simplifies considerably the mathematical analysis of the problem.

A schematic representation of the circuit system for multiplex reception is shown in Fig. 8.

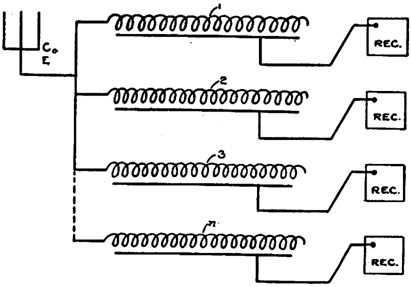

Fig. 8

Several wave conductors, n in number, are connected to the antenna which is represented by the capacity C_0 and signal e.m.f. E. The metal plates of the wave conductors are each connected to a separate receiver, the metal plate serving in this case a double purpose: a means for adjusting the wave conductors, and also as a means for coupling the wave conductors to the receivers. Other means of coupling the receivers to the wave conductors may be employed, and in that case the metal plates may be grounded if desired.

We shall designate by subscripts 1, 2, 3, · · · , n the currents and voltages on the corresponding wave conductors, and also designate the electrical constants of the wave conductor by corresponding subscripts.

We have then:

For wave conductor 1

$$\left.\begin{aligned} I_1 &= A_1 e^{k_1 x} + B_1 e^{-k_1 x}, \\ V_1 &= -\frac{k_1}{C_1 j\omega}\left(A_1 e^{k_1 x} - B_1 e^{-k_1 x}\right), \\ k_1 &= \sqrt{(L_1 j\omega + R_1) C_1 j\omega}. \end{aligned}\right\} \tag{39}$$

For wave conductor 2

$$\left.\begin{aligned} I_2 &= A_2 e^{k_2 x} + B_2 e^{-k_2 x}, \\ V_2 &= -\frac{k_2}{C_2 j\omega}\left(A_2 e^{k_2 x} - B_2 e^{-k_2 x}\right), \\ k_2 &= \sqrt{(L_2 j\omega + R_2) C_2 j\omega} \end{aligned}\right\} \tag{40}$$

Similar equations for all the other wave conductors. For the nth wave conductor, we have

$$\left.\begin{aligned} I_n &= A_n e^{k_n x} + B_n e^{-k_n x}, \\ V_n &= -\frac{k_n}{C_n j\omega}\left(A_n e^{k_n x} - B_n e^{-k_n x}\right). \end{aligned}\right\} \tag{41}$$

Each of the above equations must satisfy the condition that the current is zero at the end of the conductor, $x = l$, which gives the following:

$$\left.\begin{aligned} A_1 e^{k_1 l} + B_1 e^{-k_1 l} &= 0, \\ A_2 e^{k_2 l} + B_2 e^{-k_2 l} &= 0, \\ \cdots\cdots\cdots & \cdots \\ A_n e^{k_n l} + B_n e^{-k_n l} &= 0. \end{aligned}\right\} \tag{42}$$

From these we get the relations,

$$\left.\begin{aligned} B_1 &= -A_1 e^{2k_1 l_1}, \\ B_2 &= -A_2 e^{2k_2 l}, \\ \\ B_n &= -A_n e^{2k_n l}. \end{aligned}\right\} \tag{43}$$

At $x = 0$, we have the conditions that the voltages at each of the wave conductors are the same, and equal to the applied voltage less the antenna capacity reactance voltage drop. The current in C_0 is obviously equal to the sum of the currents in all the wave conductors at $x=0$. Hence the following conditions:

$$\left.\begin{aligned}&\frac{k_1}{C\,j\omega}(A_1-B_1)=\frac{k_2}{C_2 j\omega}(A_2-B_2)=\cdots=\frac{k_n}{C_n j\omega}(A_n-B_n),\\ &\text{and}\\ &\frac{k_1}{C_1 j\omega}(A_1-B_1)=E-\frac{1}{C_0 j\omega}(A_1+B_1+A_2+B_2+\cdots+A_n+B_n).\end{aligned}\right\}\tag{44}$$

Making use of the relation given by (43) we get

$$\begin{aligned}\frac{k_1}{C_1 j\omega}A_1e^{k_1l_1}(e^{k_1l}+e^{-k_1l})&=\frac{k_2}{C_2 j\omega}A_2e^{k_2l}(e^{k_2l}+e^{-k_2l})=\cdots\\ &=\frac{k_n}{C_n j\omega}A_ne^{k_nl}(e^{k_nl}+e^{-k_nl}),\end{aligned}\tag{45}$$

that is,

$$\left.\begin{aligned}A_2&=\frac{k_1c_2}{k_2c_1}\,\frac{e^{k_1l}}{e^{k_2l}}\,\frac{e^{k_1l}+e^{-k_1l}}{e^{k_2l}+e^{-k_2l}}\mathrm{A}_1,\\ A_3&=\frac{k_1c_3}{k_3c_1}\,\frac{e^{k_1l}}{e^{k_3l}}\,\frac{e^{k_1l}+e^{-k_1l}}{e^{k_3l}+e^{-k_3l}}A_1,\\ &\cdots\cdots\cdots\cdots\cdots\cdots\\ A_n&=\frac{k_1c_n}{k_nc_1}\,\frac{e^{k_1l}}{e^{k_nl}}\,\frac{e^{k_1l}+e^{-k_1l}}{e^{k_nl}+e^{-k_nl}}\,A_1.\end{aligned}\right\}\tag{46}$$

Introducing these values in the second equation (44), we obtain the following:

$$\begin{aligned}\frac{k_1}{C_1 j\omega}A_1e^{k_1l}(e^{k_1l}+e^{-k_1l})&=E-\frac{A_1}{C_0 j\omega}\Big\{\{(1-e^{2k_1l})\\ +\frac{k_1c_2}{k_2c_1}e^{k_1l}\frac{(e^{k_1l}+e^{-k_1l})}{e^{k_2l}+e^{-k_2l}}(1-e^{2k_2l})&+\cdots+\frac{k_1c_n}{k_nc_1}e^{k_1l}\frac{(e^{k_1l}+e^{-k_1l})}{e^{k_nl}+e^{-k_nl}}(1-e^{2k_nl})\Big\}.\end{aligned}\tag{47}$$

This simplifies to

$$\begin{aligned}\frac{k_1}{C_1 j\omega}A_1=\frac{Ee^{-k_1l}}{e^{k_1l}+e^{-k_1l}}+\frac{1}{C_0 j\omega}A_1\Big\{&\frac{e^{k_1l}-e^{-k_1l}}{e^{k_1l}+e^{-k_1l}}+\frac{k_1c_2}{k_2c_1}\,\frac{e^{k_2l}-e^{-k_2l}}{e^{k_2l}+e^{-k_2l}}+\cdots\\ &+\frac{k_1c_n}{k_nc_1}\,\frac{e^{k_nl}-e^{-k_nl}}{e^{k_nl}+e^{-k_nl}}\Big\}\end{aligned}$$

which gives

$$\left.\begin{aligned}
A_1&=\frac{Ee^{-k_1 l}}{(e^{k_1 l}+e^{-k_1 l})\left\{\dfrac{k_1}{C_1 j\omega}-\dfrac{1}{C_0 j\omega}\left(\tanh k_1 l+\dfrac{k_1 c_2}{k_2 c_1}\tanh k_2 l+\cdots+\dfrac{k_1 c_n}{k_n c_1}\tanh k_n l\right)\right\}}\\
&\text{and}\\
B_1&=\frac{-Ee^{k_1 l}}{(e^{k_1 l}+e^{-k_1 l})\left\{\dfrac{k_1}{C_1 j\omega}-\dfrac{1}{C_0 j\omega}\left(\tanh k_1 l+\dfrac{k_1 c_2}{k_2 c_1}\tanh k_2 l+\cdots+\dfrac{k_1 c_n}{k_n c_1}\tanh k_n l\right)\right\}}
\end{aligned}\right\}\qquad(48)$$

By (46) expressions for the other constants are readily obtained. Thus

$$\left.\begin{aligned}
A_2&=\frac{E\dfrac{k_1 c_2}{k_2 c_1}e^{-k_2 l}}{(e^{k_2 l}+e^{-k_2 l})\left\{\dfrac{k_1}{C_1 j\omega}-\dfrac{1}{C_0 j\omega}\left(\tanh k_1 l+\dfrac{k_1 c_2}{k_2 c_1}\tanh k_2 l+\cdots+\dfrac{k_1 c_n}{k_n c_1}\tanh k_n l\right)\right\}}\\
B_2&=\frac{-E\dfrac{k_1 c_2}{k_2 c_1}e^{k_2 l}}{(e^{k_2 l}+e^{-k_2 l})\left\{\dfrac{k_1}{C_1 j\omega}-\dfrac{1}{C_0 j\omega}\left(\tanh k_1 l+\dfrac{k_1 c_2}{k_2 c_1}\tanh k_2 l+\cdots+\dfrac{k_1 c_n}{k_n c_1}\tanh k_n l\right)\right\}}
\end{aligned}\right\}$$

and similarly for the other constants, that is

$$\begin{aligned}
A_n&=\frac{E\dfrac{k_1 c_n}{k_n c_1}e^{-k_n l}}{(e^{k_n l}+e^{-k_n l})\left\{\dfrac{k_1}{C_1 j\omega}-\dfrac{1}{C_0 j\omega}\left(\tanh k_1 l+\dfrac{k_1 c_2}{k_2 c_1}\tanh k_2 l+\cdots\right)\right\}},\\
B_n&=\frac{-E\dfrac{k_1 c_n}{k_n c_1}e^{k_n l}}{(e^{k_n l}+e^{-k_n l})\left\{\dfrac{k_1}{C_1 j\omega}-\dfrac{1}{C_0 j\omega}\left(\tanh k_1 l+\dfrac{k_1 c_2}{k_2 c_1}\tanh k_2 l+\cdots\right)\right\}}.
\end{aligned}\qquad(49)$$

The voltage on any one of the wave conductors, say the *m*th conductor, is given by the following:

$$\left.V_m=\frac{E\dfrac{k_1}{C_1 j\omega}\left\{e^{k_m(l-x)}+e^{-k_m(l-x)}\right\}}{(e^{k_m l}+e^{-k_m l})\left\{\dfrac{k_1}{C_1 j\omega}-\dfrac{1}{C_0 j\omega}\left(\tanh k_1 l+\dfrac{k_1 c_2}{k_2 c_1}\tanh k_2 l+\cdots+\dfrac{k_1 c_n}{k_n c_1}\tanh k_n l\right)\right\}},\right\}\qquad(50)$$

which may be written in this form

$$V_m=\frac{E\left\{e^{k_m(l-x)}+e^{-k_m(l-x)}\right\}}{(e^{k_ml}+e^{-k_ml})\left\{1-\frac{1}{C_0j\omega}\left(\frac{C_1j\omega}{k_1}\tanh k_1l+\frac{C_2j\omega}{k_2}\tanh k_2l+\cdots\frac{C_nj\omega}{k_n}\tanh k_nl\right)\right\}}\tag{51}$$

For $x=l$

$$V_m=\frac{2E}{(e^{k_ml}+e^{-k_ml})\left\{1-\frac{1}{C_0j\omega}\left(\frac{C_1j\omega}{k_1}\tanh k_1l+\frac{C_2j\omega}{k_2}\tanh k_2l+\cdots\frac{C_nj\omega}{k_n}\tanh k_nl\right)\right\}}\tag{52}$$

To see more clearly what little effect additional wave conductors have on the efficiency of reception on any one wave conductor, it is best to express (52) trigonometrically using the approximations of (22). For wave conductor (1), we have V_1 equal to a fraction the numerator of which is E and the denominator of which is

$$\begin{aligned}(\cos\beta_1l+j\alpha_1l\sin\beta_1l)&-\frac{1}{C_0j\omega}\sqrt{\frac{C_1}{L_1}}(\alpha_1l_1\cos\beta_1l_1+j\sin\beta_1l_1)\\&-\frac{1}{C_0j\omega}\sqrt{\frac{C_2}{L_2}}(\cos\beta_1l_1+j\alpha_1l_1\sin\beta_1l_1)\tanh k_2l-\cdots\\&-\frac{1}{C_0j\omega}\sqrt{\frac{C_n}{L_n}}(\cos\beta_1l_1+j\alpha_1l_1\sin\beta_1l_1)\tanh k_nl.\end{aligned}\tag{53}$$

If the wave conductor is adjusted for a particular signal frequency so that $\beta_1l=\pi/2$, the above reduces to

$$V_1=\frac{E}{j\alpha_1l-\frac{1}{C_0\omega}\sqrt{\frac{C_1}{L_1}}-\frac{1}{C_0\omega}\sqrt{\frac{C_2}{L_2}}\alpha_1l\tanh k_2l-\cdots\frac{1}{C_0\omega}\sqrt{\frac{C_n}{L_n}}\alpha_1l\tanh k_nl.}\tag{54}$$

All of the terms in the denominator which have the factor αl are very small in comparison with the first two terms unless $\tanh kl$ for any of the wave conductors is very large, which would be the case only when any of the other wave conductors should be in adjustment for a frequency extremely close, say 1/2 of one per cent, to the frequency of wave conductor (1).

Assuming that the signals to be received on the different wave conductors are of appreciably different frequencies, then the expressions for the voltage resonance on any one of the wave conductors reduces to the following:

$$V_m = \frac{E}{j\alpha_m l - \frac{1}{C_0\omega}\sqrt{\frac{C_m}{L_m}}},$$

$$= \frac{E\sqrt{\frac{L_m}{C_m}}}{j\frac{R_m l}{2} - \frac{1}{C_0\omega}} = -EC_0\omega\sqrt{\frac{L_m}{C_m}} \text{ approx.} \qquad (55)$$

This system offers also the possibility of operating a number of receivers on the same frequency from a single antenna.

If all the wave conductors are adjusted for resonance for the same frequency,

$$k_1 l = k_2 l = k_3 l = \cdots k_n l.$$

and (52) reduces to

$$V = \frac{2E}{(e^{kl}+e^{-kl})\left\{1 - \frac{n}{C_0 j\omega}\,\frac{Cj\omega}{k}\tanh kl\right\}}. \qquad (56)$$

Expressed in trigonometric terms using the approximations of (22), the above transforms to

$$V = \frac{E}{(\cos\beta l + j\alpha l\sin\beta l) - \frac{n}{C_0 j\omega}\sqrt{\frac{C}{L}}(\alpha l\cos\beta l + j\sin\beta l)}. \qquad (57)$$

If each of the wave conductors is adjusted for the condition

$$\cos\beta l = \frac{n}{C_0\omega}\sqrt{\frac{C}{L}}\sin\beta l,$$

which is the resonance condition for the system, the voltage on each wave conductor is given by

$$V = \frac{E}{\alpha l\left\{j\sin\beta l + j\frac{n}{C_0\omega}\sqrt{\frac{C}{L}}\cos\beta l\right\}}$$

$$= \frac{E}{j\alpha l\sin\beta l\left\{1 + \frac{n^2}{C_0^2\omega^2}\,\frac{C}{L}\right\}}. \qquad (59)$$

The term $n^2/C_0^2\omega^2\ C/L$ is small compared with the unity term, unless n is fairly large, and for a small number of receivers the resonance voltage is effected but very little by additional receivers. We may assume the following values: $1/C_0\omega=200$, and $L/C=10^8$, and for these values we have

$$
\begin{aligned}
n=10; &\quad 1+\frac{n^2}{C_0^2\omega^2}\,\frac{C}{L}=1.04\\
n=20; &\quad \text{``} \qquad =1.16\\
n=30; &\quad \text{``} \qquad =1.36\\
n=40; &\quad \text{``} \qquad =1.64\\
n=50; &\quad \text{``} \qquad =2.00.
\end{aligned}
$$

It is seen that even for fifty receivers the resonance voltage is reduced to only one half the voltage that would develop where only one receiver is connected to the antenna. If the antenna were separately tuned to that particular frequency, then $1/C_0\omega$ would be replaced by R_0 the resistance of the antenna tuning coil which would be of course very much less than $1/C_0\omega$, and in that case adding on receivers would have no appreciable effect on either amplitude of the voltages developed on the wave conductors, or on the tuning of the wave conductors.

This multiplex method should prove useful for broadcast reception in apartment houses, one large antenna serving as a supply source of the signal energy for all the receivers in the various apartments. It would be simply a matter of extending leads from the antenna to the different apartments each connecting to a wave conductor which is coupled to a receiver. In accordance with the discussion given here a small reduction in signal strength occurs only when all the receivers are operated on the same frequency, and the probability of any large numbers of receivers operating simultaneously on the same frequency is likely to occur only when receiving from a local station in which case the energy is more than sufficient, and the reduction in intensity would not be even noticed. For distant stations, if the receiving sets are operated on different frequencies, no appreciable effect is produced on either the tuning or strength of signals by the addition of other receivers connected through wave conductors.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 10 *October, 1929*

WAVE RESONANCE TUNING AND APPLICATION TO RADIO TRANSMISSION*

By
William R. Blair and Louis Cohen
(Signal Corps, War Department, Washington, D. C)

In a previous paper by one of the authors, the theory of wave resonance tuning was discussed, and methods described for the use of this principle of tuning in the reception of radio signals.

This paper has reference to methods for the utilization of wave resonance tuning in the transmission of radio signals. Several advantages are realized by this method of tuning, the most important of which are the elimination of harmonics, extreme sharpness of tuning, and a convenient and relatively simple method for multiplex transmission.

In the use of vacuum-tube transmitters, it is seldom that the transmission is effected on a single frequency, the radiated energy almost always having harmonic frequencies superimposed on the basic frequency. The listener in on broadcasting has ample evidence of it all the time. Also in communication transmitting stations there is no lack of harmonics. Theoretically it should be possible by proper care in design and operation to produce oscillations of a sine wave form of a single frequency, but practically this is seldom realized.

There are various causes which may contribute to the distortion of the wave form of the oscillations; some are inherent in the operating characteristics of the tube itself. The fact is that even with the use of master oscillators the oscillations are not altogether free from distortion. The method proposed here for the elimination of the harmonics is very simple and effective. This has been demonstrated repeatedly in the laboratory. One of the circuit arrangements used consists in interposing a wave conductor between the source of the oscillations and the antenna in the manner shown in Fig. 1. The wave conductor *A*, which consists of a solenoidal coil mounted on a movable plate *B*, is connected to the oscillating circuit *D*, and the antenna is connected

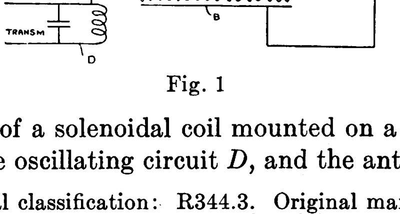

Fig. 1

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to the metal plate B. The oscillations generated in D are transmitted through the wave conductor to the antenna. The wave conductor, however, in its transmission efficiency discriminates in favor of one frequency for which it is in adjustment. Assume that the oscillations produced in circuit D are of a distorted wave form comprised of fundamental frequency and harmonics, and let the wave conductor be put in adjustment for a quarter wavelength of the fundamental frequency. Then it can be shown, the theory of which was given in the previous paper referred to, that for that frequency the wave conductor will respond resonantly and serve as an efficient agency for the transfer of the fundamental oscillations to the antenna. For other frequencies if they are even harmonics of the fundamental frequency, the wave conductor will not respond at all, and even for odd harmonics only a small fraction of the energy is transmitted to the antenna. This can be readily seen from the following considerations: for a conductor of distributed inductance and capacity, a voltage applied at one end and the other end open, the voltage distribution is given by the following equation:

$$V=\frac{E\left\{e^{k(l-x)}+e^{-k(l-x)}\right\}}{e^{kl}+e^{-kl}}, \tag{1}$$

where $k=\alpha+j\beta$ in the usual notation.

Since the energy transfer from wave conductor to the antenna is effected through electrostatic coupling, it is altogether a matter of voltage, and (1) is to be integrated to obtain an expression for the total voltage.

$$V_t=\int_0^l V dx=\frac{E}{k}\,\frac{\left[-e^{k(l-x)}+e^{-k(l-x)}\right]_0^l}{e^{kl}+e^{-kl}}, \tag{2}$$

$$=\frac{El}{kl}\,\frac{e^{kl}-e^{-kl}}{e^{kl}+e^{-kl}}.$$

For short conductors, we may put

$$e^{kl}+e^{-kl}=2\cos\beta l+2j\alpha l\sin\beta l,$$

$$e^{kl}-e^{-kl}=2\alpha l\cos\beta l+2j\sin\beta l.$$

Hence

$$V_t=\frac{El}{j\beta l}\,\frac{\{\alpha l\cos\beta l+j\sin\beta l\}}{\cos\beta l+j\alpha l\sin\beta l}. \tag{3}$$

For a frequency corresponding to a quarter wavelength, $\beta l=\pi/2$, and the above reduces

$$V_t=\frac{El}{j\pi/2\alpha l}. \tag{4}$$

Since αl is very small the voltage is great, and as a consequence efficient energy transfer to the antenna. For even harmonics $\alpha l=\pi$, 2π, 3π, etc.

$$V_t=\frac{El}{j\pi}\alpha l. \tag{5}$$

The voltage is extremely small. αl may be of the order of 0.005, and for applied voltages of the same magnitude the ratio between fundamental and even harmonic will be of the order of 25×10^{-6}, which is negligible. For odd harmonics, say $\beta l=3\pi/2$,

$$\beta l=\frac{3\pi}{2},$$

$$V_t=\frac{El}{j3\pi/2\alpha l}.$$

It would appear that the amplitude is one-third of the fundamental, but in reality it is a great deal less because $\alpha l=Rl/2\sqrt{C/L}$, and for a frequency three times greater the resistance may be several times greater. We may therefore safely assume that for the same applied voltages the response to the harmonic would be no more than about ten per cent of that to the fundamental frequency. If the amplitude of the harmonic in the oscillation source is ten per cent that of the fundamental, the effect on the antenna would be for that harmonic only one per cent that of the fundamental, also a negligible quantity. Harmonics of higher order are even more negligible in their effects on the antenna.

It follows from the above consideration that this system offers an effective means for the elimination of harmonics, no matter what the character of the wave form generated, since oscillations of only one frequency are transmitted to the antenna. This has been fully substantiated by experimental tests in the laboratory.

The sharpness of tuning, and the consequent narrowing of the transmission band emitted by the antenna when this system of tuning is employed, is evident from the discussion given in another paper to which reference has been made above.

It is possible, on the other hand, by this method of tuning to utilize the harmonics of the oscillator as separate communication channels. All that it is necessary to do is to connect additional wave conductors to the oscillator, adjusting separately each of these additional wave conductors for one of the harmonic frequencies. Each wave conductor would be provided with a separate keying device and suitable amplification, if desired. These wave conductors may operate either on separate antennae or on the same antenna.

It is observed that in the circuit arrangement shown in Fig. 1, the antenna proper is untuned. An improvement in the efficiency would result by tuning the antenna, but it is not altogether necessary.

A system in which the antenna is untuned immediately suggests the possibility of multiplex transmission, and this has actually been accomplished by the arrangement shown in Fig. 2. Each transmitter is separately associated with a wave conductor which is in adjustment for the fundamental frequency of that transmitter, and all the metal plates of the different wave conductors are connected to the antenna. It was demonstrated experimentally that there is no interaction between the different transmitters, each merely feeding energy into the antenna at its own frequency, and the current in the antenna is the sum of the currents due to the several transmitters when acting sinlgy.

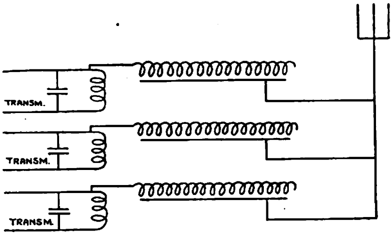

Fig. 2

The method of embodying the principle of "wave resonance" tuning described above for the elimination of harmonics and multiplex transmission is typical of many other arrangements we have used in which this principle of tuning was utilized. A more detailed discussion of the many modifications developed in the application of this method of tuning will be given at a later date.

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Proceedings of the Institute of Radio Engineers
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Discussion* on
RADIO DIRECTION-FINDING BY TRANSMISSION AND RECEPTION†

R. L. Smith-Rose

(Communicated in reply to previous discussion)

R. L. Smith-Rose:[1] I should like to take this opportunity of thanking the members of the Institute for the cordial reception given to my paper at the New York meeting on February 6th last. In particular, I wish to express my indebtedness to Dr. L. M. Hull for kindly presenting the paper on my behalf and for replying to various points which arose in the discussion.

Mr. Ballantine is undoubtedly correct in drawing my attention to the limitations of the reciprocal theorem, and I have noted the publication of his own and Mr. Carson's papers on this subject in the June number of the Proceedings. With regard to compensation of the rotating loop, some experiments were carried out on this at Gosport with little success due, I consider, to the unfavorable conditions at that station. It is expected that future beacons in Great Britain can be erected on much better sites.

Commander Taylor reviews the cause of errors in direction-finding due to downcoming waves and draws attention to the effect of the angle of incidence of these waves. The minimum distance at which "night" or "atmospheric" errors are experienced will undoubtedly depend upon the wave-frequency, since it is the relative strengths of the ground and atmospheric waves which is the deciding factor. The short ground waves will be attenuated more than the longer waves and so we should expect the errors to begin at shorter distances from the transmitter. I might also emphasize the fact that since, as pointed out by Mr. Dean, the available experimental evidence indicates that the ground conductivity in the United States is lower than in Great Britain, the ground waves will be attenuated more in the former country, and it is therefore to be expected that "night" errors in America will occur at shorter ranges than in Great Britain at the same frequencies. This probably accounts for Commander Taylor's experience of a 30-deg. error at six miles from a transmitter operating in the broadcast band of wavelengths. This is quite contrary to all our experience in England. The reception sensitivity of the Adcock aerial system will, of course, increase as the wavelength is reduced; but in connection with our work we have generally been able to obtain sufficient sensitivity with the aid of modern receivers. I am pleased to notice that Mr. Putnam has supported my contention as to the equality of errors experienced on CW, ICW, and spark transmissions. I am further gratified to note that the last section of my paper dealing with the very controversial subject of the relative advantages of the direction-finder and the rotating beacon in marine navigation has produced exactly the response I desired. Mr. Putnam has given me much useful information from his greater experience of the use of directional wireless in navigation. While being in close agreement with most of the points raised I think that there will still be room for the rotating beacon in providing an additional navigational service, particularly from the smaller type of craft which would not instal a direction-finder. At the time of writing the first rotating beacon for the use of the mercantile marine has just been put into commission at Orfordness on the East coast of England.

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* Received August 6, 1929.

† Presented before New York meeting of the Institute, February 6, 1929. Proc. I.R.E., 17, 425; March, 1929.

[1] National Physical Laboratory, Teddington, England.

I am indebted to Dr. Schelleng for his discussion of the behavior of directional errors with reversal of the direction of propagation in the presence of the earth's magnetic field. I trust that I did not give the impression in my paper that Fig. 14 and the last paragraph of section 4 on page 461 provided a proof of the instantaneous reversibility of night errors with a direction-finder and a rotating beacon. The remarks at the bottom of page 438 provide the necessary reservation to the application of the reciprocal theorem.

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Proceedings of the Institute of Radio Engineers
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Discussion* on
MICROPHONIC IMPROVEMENT IN VACUUM TUBES†

Alan C. Rockwood and Warren R. Ferris
(Research Laboratory, General Electric Company, Schenectady, New York)

M. J. Kelly:[1] I should like to ask if the filament of the UX864 is mounted with any attempt at a fixed tension.

Alan C. Rockwood:[2] Only insofar as seeing that the filament is symmetrical.

M. J. Kelly: Is it correct then that you obtain the low microphonic response described without holding to narrow limits the tension exerted on the filament by the spring at the top of the *V* of filament?

Alan C. Rockwood: Yes. The adjustment is left to the operator who adjusts the filament spring as she would for any coated filament tube with particular attention to seeing that the hook is centered on the filament. The highest disturbance occurs if an uneven filament is secured so one leg is slack. If you have a cylindrical tube high filament tension will give trouble, while with a *V* or *W* filament tube an excessively loose filament leg will cause trouble.

M. J. Kelly: Is there any evidence of a sputtering noise, somewhat similar to static and different from the microphonic noise due to variable contact between the hook of the spring and the filament?

Alan C. Rockwood: We have had no evidence of that. Such a noise might be either the result of a charge accumulating on the filament hook or of the filament sliding along the hook, but we have never found it when looking for it in this tube.

A. E. Rodd:[3] In reference to microphonic noises caused with the 227 type of tube, how about the new type with filament suspended in the center of the cathode?

Alan C. Rockwood: The microphonic characteristics of that type of 227 are no different from the old.

A. E. Rodd: The new type of cathode sleeve with the filment suspended in the center?

Alan C. Rockwood: The chief microphonic response in the 227 has been always the overall response of the whole mount assembly. Changing the interior construction of the cathode doesn't affect it. The oscillograms show the 227 has practically the lowest microphonic response of any of the small tubes with the exception of the UX864 and the 7½-watt power tubes.

A. E. Rodd: I mean the one without it.

Alan C. Rockwood: We have tried considerable quantities of both types in broadcast receivers and find no differences between the two types.

J. E. Pristas:[4] Is the 112A type much less microphonic than the 201A type? Is there any difference?

Alan C. Rockwood: Any difference in microphonics is probably a matter of coincidence. There is a certain amount of variation for both of them. On an actual percentage basis there is little difference.

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* Received by the Institute, August 7, 1929.

† Presented before New York meeting of the Institute, June 5, 1929. Proc. I.R.E., 17, 1621 September, 1929.

[1] Bell Telephone Laboratories, New York, N. Y.

[2] Research Laboratory, General Electric Co., Schenectady, N. Y.

[3] Non-member.

[4] Davega, Inc., New York City.

J. E. Pristas: Excuse me for differing on that. I have made a pretty accurate test in the laboratory; in practically all cases I found the 112A was less microphonic than the 201A.

Alan C. Rockwood: In testing several thousand tubes as detectors we haven't found much difference, although any variation is in favor of the coated filament tube. However, if you are referring to high gain laboratory amplifiers it is hard to predict what will happen with a certain type of tube. This will depend upon the layout and whether or not you have abnormally good tubes. It should be possible to pick out tubes of either type which are equally good if you have enough to choose from. The oscillograms [Figs. 6 and 7] show the extremes possible in the quarter ampere tube—with the poorer one being about the worst we have ever found in this structure.

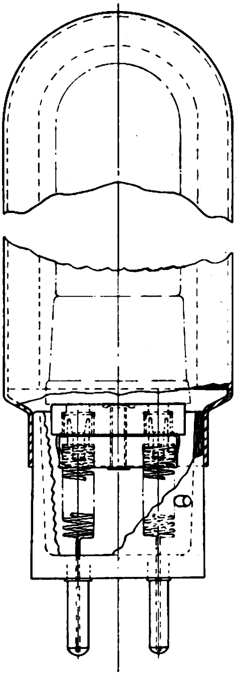

Fig. 20

Probably the best answer to your question is to say that in deciding between the UX201A and UX112A we would base a decision upon the desired electrical characteristics and not upon any differences in microphonics.

K. H. Wood:[5] I should like to ask about the 199 type tube. That is the worst offender, I believe. I wonder if any improvements were made in that type of tube.

Alan C. Rockwood: The limitation in that type of tube is the fact that economy of filament current was the primary consideration in its design and this required a small, light filament. We have to balance the microphonic response against the filament characteristics so the tube will be both usable and economical. Of course, improvements of detail have been made since the tube has first been introduced so as to better fit it to the service for which it was intended. However, we can't make the great improvement of the UX864 in this type as the use of a *V* filament implies extra end loss and loss of emission.

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[5] Non-member.

A. Hoyt Taylor:[6] I should like to ask Mr. Rockwood if there is any gain in any of the schemes for shielding the tube.

Alan C. Rockwood: An adapter was shown by an additional slide [Fig. 20] which is one of the best of the special shields which have been useful where abnormal demands have been placed on a tube. This was used for a UX199 detector placed directly under a dynamic speaker and represents about the extreme possible in a spring suspended socket with a surrounding acoustic shield. When you go to aircraft receivers, however, it has been found that with such an adapter you have so much motion of the tube as flexibility of the mounting is increased that you have trouble from change of tuning and of feedback before acoustic isolation is reached. These troubles limit the amount you can go in acoustic shielding.

A. Hoyt Taylor: I should like to point out that while discussion by the authors of the paper has confined itself largely to the interest of broadcast receivers and airplane receivers, there are two other fields in which the suppression of microphonic disturbance of tubes is of very great importance.

One of these is the high-frequency receiver, particularly the receiver for extremely high frequency. It is well known by all those who have had experience with such receivers that the microphonic troubles in tubes increase very rapidly with the frequency as far as their disastrous effects upon reception is concerned. In other words, tubes that will be perfectly all right, say at 1000 kc, are no good at all at 20,000 kc because of the extreme microphonic response and the disturbance of the circuit condition since we are primarily interested in CW reception where one of the tubes at least is oscillating.

The development of suitable tubes for very high frequency with a far higher degree of improvement of the microphonic properties than perhaps required even by aircraft work is therefore of great practical interest for those engaged in the development of such receivers.

There is one other field where the reduction of microphonic noises is also of very great interest and importance. That is in the power tube field. I have recently had some experience with a transmitter or several transmitters which used a master circuit followed by a balanced power amplifier circuit in working at frequency between 14,000 and 30,000 kc.

The circuit did not use crystal control. The small disturbances which in this case reflect on the frequency rather than volume or output would occur. But with the self oscillating master circuit of a special type (in fact it was the oscillator known as the Gunn oscillator) we found that while the master oscillator alone gave an excellent quality signal of very small frequency ripple, yet when we put a 750-watt balanced power amplifier outfit on top of this, we got a tremendous ripple in frequency.

Tracing this down we found this was entirely due to a microphonic disturbance originating (from the plate principally) in the power amplifier tubes. It was due to small variations in the effective capacity of the circuit which was loaded on to the power amplifier reacting back and influencing the frequency of the master circuit.

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[6] Naval Research Laboratory, Bellevue, Anacostia, D. C.

We do not overlook any opportunities to learn more about the frequency control, and the crystal is not the only way of getting pretty constant frequencies. I have recently had the opportunity of examining a rather new type of transmitter tube in which an attempt was made to suppress as far as possible the microphonic disturbance in the shielded transmitter tube for the purpose of using it in such special circuits which did not use piezo-electric or equivalent control. There appear to be many possibilities of great improvement in transmitter tubes as well as receiver tubes in this matter of microphonic disturbances.

I think possibly one of the most important fields for it is in the aviation field. I spent four years in that work developing aviation radio, and ten years ago we were experimenting with a sort of hectocomb unit, a unit of six tubes consisting of three stages of radio frequency with a detector and two stages of audio. We copied this outfit more or less completely from a British wartime airplane receiver equipment, and we had to find a tube that had the proper constants to go with these circuits. We had to use a tube that is no longer produced in this country. I won't name it, but I think many of you can guess who made it at the time. They were very microphonic, but very non-uniform and I can remember on many an airplane flight we would test as many as twenty-four tubes before we got a set of six tubes we could wear with the earphones with the helmets on, without the microphonic disturbance pretty nearly driving us crazy. That shows the importance of reducing microphonic disturbance from the standpoint of aviation.

That is particularly true in dealing with receivers of high amplification as we were then, because this particular work centered about radio compass work that was done preparatory to the flight of the NC-4 to Europe across the Atlantic.

I feel that if anyone can improve very materially the microphonic properties of tubes for aviation service, they will have conferred a great benefit upon humanity.

J. E. Pristas: In reference to the screen-grid tube, how does this type compare with the three-element tubes?

Alan C. Rockwood: The microphonic response of a screen-grid tube is about the same as any three-element tube of the same general type. The three-electrode detector gives the chief trouble because of the amplification following it. Similarly, the microphonic disturbance of the screen-grid tube is largely a question of how much gain you are going to have between the detector and the output tube.

Lee Sutherlin[7]**:** I should like to make one statement with respect to the microphonic study which it seems to me may bear on two or three questions asked here. In making the study of a lot of tubes, I found in general that the microphonic noise of a tube of a given current and voltage with a ribbon type filament is less. Of course, that would be a coated filament.

That may be due to the fact that it is a ribbon and vibrates only in one plane, and, of course, it may be due to the difference in materials. In many cases I have found that difference. It seems to me this bears on some of the questions that were asked.

Alan C. Rockwood: That is one of the things we have never run into, although it is logical to expect this in the heavier current filaments. For the lower current filaments it is hard to say whether any difference, as compared to tungsten filaments, is due to the cross-section or the greater mass in a single leg of a coated filament having the same rating.

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[7] Westinghouse Electric and Manufacturing Co., East Pittsburgh, Penna.

BOOK REVIEW

Radio Encyclopedia†—Encyclopedic dictionary of radioelectricity, by Michel Adam, engineer, E. S. E. One volume of VI—355 pages,[1] in quarto, with 1480 illustrations, 150 abbreviations and symbols, two plates of schematic symbols, 19 tables and five maps, published by Etienne Chiron, 40 rue de Seine, Paris, 1928. Price 61 French francs.

Although there are already several dictionaries and vocabularies of radioelectricity, in French and other languages, Mr. Adam's work is opportune. He condenses, in a practical and easily accessible form, technical, technological, and philological knowledge. The encyclopedia is in two parts:

I. Alphabetical index of abbreviations and symbols:

II. Encyclopedia of technical terms.

The encyclopedia, as it has been rightly termed, contains the names, definition, explanation, and an analytical description of all terms used in radioelectricity. It gives related terms of electricity and elementary mechanics and groups the different units of French, English and international measure used frequently in this subject. Moreover, the English and German translation of each term is given, this being absolutely necessary for all those who wish to follow the constant progress of the French and English scientific literature of radioelectricity. Numerous drawings and illustrations complete and make more interesting the literary and technical documentation, while facilitating the investigations of the novice as well as the works of the professional. The special alphabetical index contains all the French, English and German, and sometimes even Italian and Spanish abbreviations for radioelectricity with the exception of purely telegraph abbreviations, which interest only radiotelegraph operators and are treated in special works.

Thus, the "Encyclopedia of Radio," which answers now a real need of French, scientific literature, will certainly be welcomed favorably by Administrations, as well as by the public, both technicians and laymen.

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† Translated from *Journal Télégraphique*, **53**, No. 5, 113–114; May, 1929.

[1] Translator's Note: VI—355 refers to six introductory and 355 text pages.

BOOKS RECEIVED

The Physical Principles of Wireless, by J. A. Ratcliffe, Fellow and lecturer of Sidney Sussex College, Cambridge University Demonstrator in the Cavendish Laboratory. Published by E. P. Dutton and Company, Inc., 1929. 104 pages, illustrated, 4 x 6½ inches, cloth binding. Price, $1.15. Contents: oscillatory circuits; valves; wireless transmitters; reception of wireless signals; wireless telephony; amplifiers; miscellaneous.

Spectra, by R. C. Johnson, lecturer in physics in the University of London, Kings College, and O. W. Richardson, Yarrow research professor of the Royal Society. Published by E. P. Dutton and Company, Inc., 104 pages, illustrated, 4 x 6½ inches. Contents: the quantum theory; line spectra; band spectra; spectroscopy; appendix.

ABC of Television, by Raymond Francis Yates, formerly editor of *Popular Radio.* Published by The Norman W. Henley Publishing Company, 1929. 210 pages, illustrated, 5½ x 8½ inches, cloth binding. Price, $3.00. Contents: television—the new conquest of space; television systems; telegraphing pictures;

photoelectric cells—the eyes of television; amplifying pictures; the agile neon lamp; selenium cells; the problem of scanning; synchronizing television; transmitting television at home; how to make a television receiver.

Mechanical Characteristics of Transmission Lines, by Seiichiro Noda of the Ryojun College of Engineering, Port Arthur, South Manchuria. Published as Vol. II, No. 2 of the Memoirs of that institution. 155 pages, illustrated, charts and drawings, 7½ x 10½ inches, paper cover. Contents: general solution, law of geometric progression, breaking of a wire in a span of continuous wire-spans, partial change in load on continuous wire-spans, failure of a suspension insulator in continuous wire-spans.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 10 *October, 1929*

MONTHLY LIST OF REFERENCES TO CURRENT RADIO LITERATURE

THIS is a monthly list of references prepared by the Bureau of Standards and is intended to cover the more important papers of interest to professional radio engineers which have recently appeared in periodicals, books, etc. The number at the left of each reference classifies the reference by subject, in accordance with the scheme presented in "A Decimal Classification of Radio Subjects—An Extension of the Dewey System," Bureau of Standards Circular No. 138, a copy of which may be obtained for 10 cents from the Superintendent of Documents, Government Printing Office, Washington, D. C. The various articles listed below are not obtainable from the Government. The periodicals can be secured from their publishers and can be consulted at large public libraries.

R100. RADIO PRINCIPLES

R100 Morecroft, J. H. Elements of radio communication (book). John Wiley and Sons, Inc., 1929. New York, N. Y.

(A simple text book of radio introductory to the more comprehensive text "Principles of Radio Communication" by the same author.)

R112.1 Strutt, M. J. O. Strahlung von Antennen unter dem Einfluss der Erdbodeneigenschaften; (a) elektrische Antennen, (b) magnetische Antennen. (Radiation from antennas and effects of the earth's influence). *Annalen der Physik*, pp. 721–772; 1929.

(A thorough discussion of the radiation characteristics of both horizontal and vertical antennas especially of the doublet type.)

R113 Yokoyama, E. and Nakai, T. East-West and North-South attenuations of long radio waves on the Pacific. PROC. I. R. E., **17**, pp. 1240–1247; July, 1929.

(A comparative study of the observed values of field intensity of low-frequency high-powered stations in the Pacific region seems to show a much greater east-west than north-south attenuation during daylight hours in fairly high latitudes. A comparison of observed values with values calculated by various transmission formulas indicates the need for the inclusion in the formulas of a factor depending on direction and latitude.)

R113.4 Ponte, M. and Rocard, Y. Sur la couche ionisee de la haute atmosphere. (On the ionized layer of the upper atmosphere.) *L'Onde Electrique*, **8**, pp. 179–191; May, 1929.

(In problems concerning the conductivity of the Heaviside layer the mean free path of the electrons is a relevant quantity. An analysis based on the kinetic theory in which the forces of interaction between the molecules and the electrons are considered shows value of the mean free path to be from 40 to 160 times as small as value obtained by the elementary theory.)

R113.5 Hulburt, E. O. On the ultra-violet light theory of aurorae and magnetic storms. *Phys. Review*, **34**, pp. 344–351; July 15, 1929.

(Certain details of the ultra-violet light theory of auroras and magnetic storms are developed and experimental facts which are in accord with the theory are cited.)

R113.5 Wymore, I. J. The relation of radio propagation to disturbances in terrestrial magnetism. PROC. I. R. E. **17**, pp. 1206–1213; July, 1929.

(The results of a study of the relationship between radio reception and the changes in the earth's magnetism show that for daylight reception over great distances there is an increase in the intensity of received signal which reaches its maximum in from one to two days and disappears in from four to five days after a magnetic storm.)

R113.6 Pedersen, P. O. Long delayed radio echoes. *Nature* (London), **124**, p. 164; July 27, 1929.

(Article taken from paper "Wireless Echoes of Long Delay" communicated to the Physical Section of Danish Royal Society. Shows mathematically that radio echoes occurring after 10 seconds cannot be due to propagation of waves within the earth's atmosphere, that echoes occurring after intervals up to 30 seconds are due to propagation along or reflection from Störmer bands as explained in *Nature* (122, p. 681, 1928), that echoes after several minutes must be from outside the space in which the earth's magnetic field exerts appreciable effect. Transmissions at various wavelengths are also treated.)

R113.6 Discussion on "The attenuation of wireless waves over land." *Jour.* I. E. E., **67**, p. 931; July, 1929.

(Discussion between C. R. Englund and R. H. Barfield of certain applications of optical theory in Barfield's paper, "The attenuation of wireless waves over land.")

R113.6 The effect of the earth on short wave radiation from vertical and horizontal aerials (editorial). *Experimental Wireless and W. Engr.* (London), **6**, pp. 351–352; July, 1929.

(Results obtained by M. J. O. Strutt and given in a paper in the *Annalen der Physik*, p. 721, 1929 are discussed.)

R113.7 Burstyn, W. Die unmittelbare Messungen von Entfernungen durch elektrische Wellen. (Direct measurements of distances by means of electric waves). *Zeits. fur Hochfrequenztechnik,* **33**, pp. 181–183; May, 1929.

(Formulas and curves showing the relationship between intensity of reception and distance from the transmitting set are given with a discussion of their accuracy. This method of measuring distances is believed invaluable for ships in foggy weather.)

R114 Schindelhauer, F. Elektromagnetische Störungen. (Electromagnetic disturbances). *Elektrische-Nachrichten Technik,* **6**, pp. 231–236; June, 1929.

(A study of electromagnetic disturbances in the atmosphere and results of experiments carried on at the Meteorological-Magnetic Observatory at Potsdam, Germany.)

R114 Paoloni, D. B. and Ilardi, G. P. Compte rendu des observations radioatmospheriques faites pendant l'annee 1927. (An account of radio atmospheric observations made in 1927). *L'Onde Electrique,* **8**, pp. 222–226; May, 1929.

(Atmospherics are classed according to the Paoloni radio atmospheric scale. Observations for each class were taken daily during 1927 by an aural system. Observations are summarized in curves and general statements.)

R114 Dean, S. W. Correlation of directional observations of atmospherics with weather phenomena. Proc. I. R. E., **17**, pp. 1185–91; July, 1929.

(Description of the results obtained with a cathode-ray direction finder used by the A. T. and T. Co. at Houlton, Me. Direction of storms was determined with a considerable degree of accuracy.)

R114 Harper, A. E. Some measurements on the directional distribution of static. Proc. I. R. E., **17**, pp. 1214–1224; July, 1929.

(The utility of directional data on static is shown, and two types of apparatus for directional investigation are compared. A method which gives direction of individual crashes is found superior to the integrating methods. Distribution of thunderstorms over the world is discussed. Probable geographic locations are assigned to disturbances.)

R125.6 Stenzel, H. Uber die Richtcharakteristik von einer Ebene angeordneten Strahlern. (The directional characteristic of a polarized radiator). *Elektrische-Nachrichten-Technik,* **6**, pp. 165–181; May, 1929.

(Complete technical discussion of directional characteristics of polarized radio waves.)

R132 von Ardenne, M. Die aperiodische Verstärkung von Rundfunkwellen. (Aperiodic amplification at broadcast frequencies.) *Zeits. fur Hochfrequenztechnik,* **33**, pp. 166–175; May, 1929.

(The problem of aperiodic cascade amplification is thoroughly discussed. Several methods of a-f and r-f amplification are given, and are discussed theoretically.)

R133 Kirschstein, F. Zur Theorie der rückgekoppelten Röhrensenders. (Theory of back-coupled tube transmitters). *Zeits. f. Hochfrequenztechnik,* **33**, pp. 201–211; June, 1929.

(Graphical and integral calculus methods are used to explain the functioning of feed-back tube generators).

R134 Ballantine, S. Detection at high signal voltages. Part 1. Plate rectification with the high-vacuum triode. Proc. I. R. E., **17**, pp. 1153–1177; July, 1929.

(Article contains a descriptive theory, a mathematical theory for small modulation and large carrier voltages, an apparatus setup and results, and practical applications of reception at high voltages.)

R138 MacNabb, V. C. The production of emission from oxide-coated filaments. *Jour.* Optical Soc. of America and Rev. of Scientific Instruments, **19**, pp. 33–41; July, 1929.

(Results of investigation of various methods for producing oxide filaments of barium and strontium in an emitting condition.)

R140 Watanabe, Y. Theorie des gekoppelten Schwingungskreises mit Selbsterregung. (Theory of self-excited inductively coupled oscillatory circuits). *Elektrische-Nachrichten Technik,* **6**, pp. 194–210; May, 1929.

(Theory of oscillatory circuits from a mathematical viewpoint.)

R145.3 Grover, F. W. A comparison of the formulas for the calculation of the inductance of coils and spirals wound with wire of large cross section. Bureau of Standards Journal of Research, July, 1929. Research Paper No. 90. Obtainable from the Superintendent of Documents, Government Printing Office, Washington, D. C.

(Two methods for calculating the inductance of coils of wire having a relatively large cross section are given.)

R170 Dieckmann, M. Beitrag zur Beschreibung des Interferenzgebietes in der Nahe von Empfangs Antennen. (Discussion on interference regions in the neighborhood of receiving antennas). *Zeits. fur Hochfrequenztechnik,* **33**, pp. 161–166; May, 1929.

(By making simplified assumptions the interference regions in the vicinity of receiving antennas are discussed, and pictured by models and drawings. Instantaneous as well as average values of field intensity are given.)

R200. RADIO MEASUREMENTS AND STANDARDIZATION

R201.5 Ferguson, J. G. Shielding in high frequency measurements. *Bell Sys. Tech. Jour.,* **8**, pp. 560–590; July, 1929.

(The purpose and usefulness of shielding for high-frequency measurements are outlined. General principles of electrostatic shielding are developed as applied to simple impedances and to networks of impedances, particularly to bridge networks. Partial applications of these principles to the shielding of adjustable impedances and in the construction of actual bridge circuits are described.)

R201.6 Oatley, B. A. A low power audio-frequency current supply for general laboratory use. *Jour. Scientific Instruments* (London), **6**, p. 217; July, 1929.

(A description is given of apparatus for providing alternating current of constant frequency and good wave form for use in bridge measurements. The apparatus derives its power from the 100-volt d-c line. Arrangements are provided whereby several observers can use the apparatus simultaneously without mutual interference.)

R202 Kallmann, H. E. Rechteckige Verformung von Resonanzkurven nach einem neuen Prinzip und ihre Anwendung beim Empfang sehr kurzer Wellen. (A new principle in the interpretation of resonance curves and its application to short-wave reception). *Zeits. fur Hochfrequenztechnik*, **33**, pp. 212–223; June, 1929.

(The sensitivity and selectivity of short-wave (of the order of 30 cm) receiving sets is thoroughly discussed; also a new method for improving both.)

R210 Giebe, E. and Scheibe, A. Internationale Vergleichung von Frequenznormalen fur elektrische Schwingungen. (International comparisons of frequency standards). *Zeits. fur Hochfrequenztechnik*, **33**, pp. 176–180; May, 1929.

(Piezo oscillator from the Bureau of Standards and piezo resonators from Germany were tested by the laboratories in Germany, Italy, France, England, and United States and found to vary within ±1 part in 25,000.)

R210 Bogardus, H. L. and Manning, C. T. The routine measurement of the operating frequencies of broadcast stations. Proc. I. R. E., **17**, pp. 1225–39; July, 1929.

(The method of making "zero beat" measurements of the operating frequencies of broadcasting stations in the Second Radio District is described, showing the method of comparing received signal from a broadcasting station with a signal of known frequency obtained from a 10-kc multivibrator controlled by a 90-kc quartz crystal.)

R210 Marrison, W. A. A high precision standard of frequency. Proc. I. R. E., **17**, pp. 1103–1122; July, 1929. Also in *Bell Sys. Tech. Jour.*, **8**, pp. 493–514; July, 1929.

(A new standard of frequency is described in which 100,000-cycle quartz crystal-controlled oscillators of very high constancy are employed. These are inter-checked automatically and continuously with a precision of about one part in one hundred million. They are checked daily in terms of radio time signals by the usual method employing a clock controlled by current maintained at a submultiple of the crystal frequency. Specially shaped crystals are used which have been adjusted to have temperature coefficients less than 0.0001 per cent per degree C.)

R214 Lack, F. R. Observations on modes of vibration and temperature coefficients of quartz crystal plates. Proc. I. R. E., **17**, pp. 1123–1141; July, 1929. Also in *Bell Sys. Tech. Jour.*, **8**, pp. 515–535; July, 1929.

(The characteristics of piezo-electric quartz crystal plates of the perpendicular or Curie cut are compared with 30-degree or parallel-cut plates with reference to the type of vibration of the most active modes, the frequency of these modes as a function of the dimensions, and the magnitude and sign of the temperature coefficients of these frequencies. The relation of various dimensional cuts to the temperature coefficient is discussed. The analysis offers an explanation of the low temperature coefficients which can be produced by a proper choice of the dimensional ratios.)

R270 Austin, L. W. Experiments in recording radio signal intensity. Proc. I. R. E., **17**, pp. 1192–1205; July, 1929.

(The method of recording intensity of long-wave radio signals used at the Bureau of Standards and some of the results obtained are given. Variability of wave propagation in regard to field intensity and angle of incidence is shown in curves. An apparent connection between night signal variation and magnetic storms is shown. The downcoming waves seem to be reflected from rapidly changing masses of ionized gas.)

R275 Lamb, J. J. The modulometer. *QST*, **13**, pp. 8–15; Aug., 1929.

(A simple device for measuring the percentage of modulation and generally checking the performance of the phone transmitter is described. Modulometer is essentially the adaptation of the electron-tube peak voltmeter for modulation measurements as previously outlined by C. B. Jolliffe.)

R300. Radio Apparatus and Equipment

R320 Meissner, A. Transmitting antennas for broadcasting. Proc. I. R. E., **17**, pp. 1178–1184; July, 1929.

(Fundamental requirements for best broadcast aerials are stated. Aerials having a height of λ/2 are found to produce less upward and greater horizontal radiation than aerials having a height of λ/4, and consequently fading due to interference of indirect with ground waves is greatly reduced. Comparative efficiency of vertical and horizontal transmitting antennas at different frequencies, is presented on basis of limited tests.)

R330.4 Bedeau, F. and DeMare, J. Etude de la methode de Beatty pour la mesure de l'amplification d'un etage a resonance. (Study on the Beatty method of measuring the amplification of a resonant stage). *L'Onde Electrique*, **8**, pp. 210–211; May, 1929.

(The method of Beatty (*Experimental Wireless and W. Engr.*, **5**, January, 1928) for calculating the overall amplification of a resonant stage from the tube design. Circuit constants are discussed. The method involving vectorial analysis is shown useful in design.)

R331 Bartels, H. Uber die Maximalleistungen von Schutznetzleistungrohren. (On the maximum output of screen-grid tubes). *Elektrische-Nachrichten Technik*, **6**, pp. 188–193; May, 1929.

(A comparison of screen-grid with ordinary tubes, bringing out the advantages of the former.)

R342.15 Dijksterhuis, P. R. and Groeneveld, Y. B. F. J. Low-frequency amplification with transformers. *Experimental Wireless and W. Engr.* (London), **6**, pp. 374–379; July, 1929.

(It is claimed that the problems of low-frequency amplification by transformers are met in the design of the Philip's transformer (Dutch). A nickel-iron alloy core is employed to give a high primary self-inductance independent of amplitude of the voltages to be amplified. A high resistance nickel-alloy wire in the secondary coil is used to prevent a a leakage resonance peak in the characteristic curve. A special silver alloy is used in the primary winding to give low resistance and high mechanical strength.)

R344.3 Watanabe, Y. Uber den Zwischenkreisrohrensender mit stark gedampften Sekundarkreis. (On the filter circuit of tube transmitting sets with strongly damped secondary circuit). *Elektrische-Nachrichten Technik*, **6**, pp. 244–248; June, 1929.

(Mathematical analysis of this type of transmitting set giving equations and results of experimental tests.)

R360 Von Ardenne, M. Receiver with aperiodic high-frequency amplification. *Experimental Wireless and Wireless Engr.*, **6**, pp. 369–373; July, 1929.

(Loewe double triodes designed to give a high aperiodic radio-frequency amplification and a very low audio-frequency amplification are used in cascade in a long range receiver. The triodes have a high input resistance and permit great selectivity in the tuned stages. Intercoupling is prevented through shielding and the use of chokes and by-pass condensers.)

R376.3 Greaves, F. F., Kranz, F. W., Crozier, W. D. The Kyle condenser loud speaker. Proc. I. R. E., **17**, pp. 1142–1152; July, 1929.

(Describes a loud speaker developed on the Kyle principle of construction with a view to providing comparatively large actuating force and also sufficient amplitude of vibration to permit efficient and faithful reproduction.)

R376.3 Clarke, H. M. The moving coil loudspeaker. *Experimental Wireless and Wireless Engr.* (London), **6**, pp. 380–384; July, 1929.

(A compensation winding, reducing hysteresis, eddy currents and self induction is offered as the solution of the problem of response control of the moving coil loud speaker. By this winding the impedance of the moving coil is made practically independent of frequency, and by the supplementary use of shunting filters the natural resonance peaks of the speaker may be smoothed to give a desirable response curve.)

R386 Farrar, C. L. Band pass tuning. *Radio Engineering*, **9**, pp. 27–30; July, 1929.

(Reduction of sideband cutting in r-f amplifiers by use of band pass filters.)

R500. APPLICATIONS OF RADIO

R512 Ludenia, W. Reichweitenversuche mit Zentimeterwellen. (Distance range tests with centimeter waves). *Elektrische-Nachrichten Technik,* **6**, pp. 248–249; June, 1929.

(A short discussion of the use of wavelengths from 10 to 100 cm. Their application is believed most practical where sharp narrow beams are necessary, e.g., for fog signals and lighthouse stations.)

R526.1 Diamond, H. and Dunmore, F. W. A course-shift indicator for the double-modulation type radiobeacon. Bureau of Standards Journal of Research, July, 1929. Research Paper No. 77. Obtainable from Superintendent of Documents, Government Printing Office, Washington, D. C.

(To further increase reliability of the visual direction beacon system developed by the Bureau of Standards, a course-shift indicating instrument is described which is primarily for station use and indicates to the operator whether the course remains unvarying during operation and also aids in checking of beacon calibration.)

R800. NON-RADIO SUBJECTS

621.313.7 Demontvignier, M. Les redresseurs a oxyde de cuivre. (Copper-oxide rectifiers). *L'Onde Electrique,* **8**, pp. 192–209; May, 1929.

(A contact of copper and copper-oxide possesses rectifying properties. The use of these properties in commercial a-c rectifiers is described. An analysis with theoretical curves and actual oscillograms showing the characteristics of the input and output is given. The application of the rectifiers to radio circuits is summarized.)

621.385 Fuhrer, R. Energieverhaltnisse in Ortsnetzen mit Selbstanschlussbetrieb. (Energy relations on local automatic telephone networks). *Elektrische-Nachrichten Technik,* **6**, pp. 213–230; June, 1929.

(Extensive discussion of the energy relations in automatic telephone networks with eference to the new Munich installation. Bibliography is given.)

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CONTRIBUTORS TO THIS ISSUE

Blair, William R.: Born November 7, 1874 in County Derry, Ireland. Came to America in 1884. Graduated Kansas State Normal School, 1895. Received B. S. degree, University of Chicago, 1904; Ph D. degree, 1906. Entered govt. service, 1906; research director in charge upper air research, U.S. Weather Bureau, 1907; Physical Laboratory, 1910; executive office charge Mt. Weather Obs., Va., 1912—1914; professor meteorology, U. S. Weather Bureau since 1915; com. maj. N. A., 1917, and assigned to Aviation Corps with A. E. F. in France; transferred to signal corps, 1918; lt. col., 1919; maj. regular army, 1921; chief of research and engineering division, office of chief signal officer, Washington, since 1927; member technical subcommittee, Inter. Commn. for Aerial Navigation, Inter. Wireless Commn., Paris, 1919; meteorologist for world flight, 1924; member Inter. Radio Conf., Washington, 1927. Non-member of the Institute of Radio Engineers.

Breit, G.: See PROCEEDINGS for September, 1929.

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Cohen, Louis: Born December 16, 1876 at Kiev, Russia. Received B. Sc. degree, Armour Institute of Technology, 1901; studied University of Chicago, 1902; Ph.D. degree, Columbia University, 1905. With scientific staff, Bureau of Standards, 1905—1909; Electrical Signalling Co., 1909—1912; consulting practice since 1913; professor electrical engineering, George Washington University, 1916.; consulting engineer, War Dept., 1920—1924; U. S. delegate Prov. Tech. Com., Inter. Conf. on Elec. Commn., Paris, France, 1921; member advisory tech. board., Conference on Limitation of Armaments, Washington, 1921—1922. Fellow, Institute of Radio Engineers, and several other societies.

Everitt, William Littell: Born April 14, 1900 at Baltimore, Md. Received E. E. degree, Cornell University, 1922; M. S., University of Michigan, 1926. Instructor in electrical engineering, Cornell University, 1920—1922; engineer in charge of laboratory and P. B. X. development, North Electric Mfg. Co., Galion Ohio, 1922—1924; instructor in electrical engineering, University of Michigan, 1924—1926; assistant professor of electrical engineering in charge of communication courses, Ohio State University, 1926—1929, associate professor, 1929. Department of Development and Research, American Telephone and Telegraph Co., summers 1925—1929, inclusive. Associate member, Institute of Radio Engineers, 1925.

Hafstad, L. R.: See PROCEEDINGS for September, 1929.

Harris Sylvan: See PROCEEDINGS for March, 1929.

Kusunose, Yuziro: Born January 12, 1898 at Tokyo, Japan. Graduated Tokyo Imperial University, 1922; radio engineer, Electrotechnical Laboratory, Ministry of Communications, 1922 to date. Associate member, Institute of Radio Engineers, 1925.

Lange, Edward H.: See PROCEEDINGS for June, 1929.

Myers, John A.: Born March 29, 1905 at Baltimore, Maryland. Received B. S. degree, Johns Hopkins University, 1927. Graduate student in electrical engineering, Johns Hopkins University, 1927 to date. With Westinghouse Electric and Manufacturing Co. during several past summers.

Pedersen, P. O.: Born June 19, 1874 in Denmark. Graduated in civil engineering, Royal Technical College, Copenhagen, 1897. Awarded the gold medal of the Royal Danish Society of Sciences. Assistant professor, Royal Technical College, 1909; professor in telegraphy, telephony, and radio, Royal Technical College, 1912 to date; principal of the college, 1922 to date. Member, Royal Danish Society of Sciences, 1917; member, State Control Board for Licensed Telephone Companies of Denmark, 1917, and president since 1920; president, Danish section of the International Electric Committee, 1926 to date; awarded the H. C. Orsted medal, 1927. Received Ph. D. degree, University of Copenhagen, 1929. Fellow, Institute of Radio Engineers, 1915, and member of several other national and international societies.

Sreenivasan, K.: Born 1899. Graduated in 1920. Apprentice in the hydroelectric department of the Mysore government, 1920—1922. Studied electrical engineering at the Indian Institute of Science, Bangalore, 1922—1924. Radio-frequency and field strength measurements, 1924 to date. Awarded the Associateship of the Indian Institute of Science, 1927. Non-member of the Institute of Radio Engineers.

Turner, H. M.: Born July 20, 1882 at Hillsboro, Illinois. Received B. S. degree, University of Illinois, 1910; M. S. degree, 1915; assistant instructor in electrical engineering, 1910—1912; instructor, University of Minnesota, 1912—1918; assistant professor, Yale University, 1918—1926; associate professor, 1926 to date. Frequent contributor to leading engineering journals. Associate member, Institute of Radio Engineers, 1914; Member, 1920; member of several other societies.

Tuve, M. A.: See PROCEEDINGS for September, 1929.

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Institute of Radio Engineers

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CLEVELAND SECTION
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LOS ANGELES SECTION
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NEW YORK MEETING
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PITTSBURGH SECTION
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SAN FRANCISCO SECTION
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TORONTO SECTION
Toronto, Ont., Canada, November 13, 1929

WASHINGTON SECTION
Washington, D. C., November 25, 1929

Eastern Great Lakes District Convention

November 18-19, 1929, Rochester, N. Y.

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STUART BALLANTINE G. W. PICKARD
RALPH BATCHER L. E. WHITTEMORE
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  1. Introduction, by ALFRED N. GOLDSMITH . . . 1940
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  3. The National Broadcasting Company—A Technical Organization for Broadcasting, by JULIUS WEINBERGER . . . 1969
  4. Speech Input Equipment, by D. G. LITTLE . . . 1986
  5. Wire Line Systems for National Broadcasting, by A. B. CLARK . 1998
  6. Radio Broadcasting Regulation and Legislation, by J. H. DELLINGER . . . 2006
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The Penetration of Rock by Electromagnetic Waves and Audio Frequencies, by A. S. EVE, D. A. KEYS, AND F. W. LEE . . . 2072
A Vacuum-Tube Voltage Regulator for Large Power Units, by L. C. VERMAN AND H. J. REICH . . . 2075
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The Institute of Radio Engineers

GENERAL INFORMATION

The PROCEEDINGS of the Institute is published monthly and contains papers and discussions thereon submitted for publication or for presentation before meetings of the Institute or its Sections. Payment of the annual dues by a member entitles him to one copy of each number of the PROCEEDINGS issued during the period of his membership.

Subscription rates to the PROCEEDINGS for the current year are received from non-members at the rate of $1.00 per copy or $10.00 per year. To foreign countries the rates are $1.10 per copy or $11.00 per year.

Back issues are available in unbound form for the years 1918, 1920, 1921, 1922, and 1926 at $9.00 per volume (six issues) or $1.50 per single issue. Single copies for the year 1928 are available at $1.00 per issue. For the years 1913, 1914, 1915, 1916, 1917, 1918, 1924, and 1925 miscellaneous copies (incomplete unbound volumes) can be purchased for $1.50 each; for 1927 at $1.00 each. The Secretary of the Institute should be addressed for a list of these.

Discount of twenty-five per cent on all unbound volumes or copies is allowed to members of the Institute, libraries, booksellers, and subscription agencies.

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Year Books for 1926, 1927, and 1928, containing general information, the Constitution and By-Laws, catalog of membership etc., are priced at seventy-five cents per copy per year.

Contributors to the PROCEEDINGS are referred to the following page for suggestions as to approved methods of preparing manuscripts for publication in the PROCEEDINGS.

Advertising rates for the PROCEEDINGS will be supplied by the Institute's Advertising Department, Room 802, 33 West 39th Street, New York, N. Y.

Changes of address to affect a particular issue must be received at the Institute office not later than the 15th of the month preceding date of issue. That is, a change in mailing address to be effective with the October issue of the PROCEEDINGS must be received by not later than September 15th. Members of the Institute are requested to advise the Secretary of any change in their business connection or title irrespective of change in their mailing address, for the purpose of keeping the Year Book membership catalog up to date.

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It is understood that the statements and opinions given in the PROCEEDINGS are views of the individual members to whom they are credited, and are not binding on the membership of the Institute as a whole.

All correspondence should be addressed to the Institute of Radio Engineers, 33 West 39th Street, New York, N. Y., U. S. A.

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SUGGESTIONS FOR CONTRIBUTORS TO THE PROCEEDINGS

Preparation of Paper

Form—Manuscripts may be submitted by member and non-member contributors from any country. To be acceptable for publication manuscripts should be in English, in final form for publication, and accompanied by a summary of from 100 to 300 words. Papers should be typed double space with consecutive numbering of pages. Footnote references should be consecutively numbered and should appear at the foot of their respective pages. Each reference should contain author's name, title of article, name of journal, volume, page, month, and year. Generally, the sequence of presentation should be as follows: statement of problem; review of the subject in which the scope, object, and conclusions of previous investigations in the same field are covered; main body describing the apparatus, experiments, theoretical work, and results used in reaching the conclusions and their relation to present theory and practice; bibliography. The above pertains to the usual type of paper. To whatever type a contribution may belong, a close conformity to the spirit of these suggestions is recommended.

Illustrations—Use only jet black ink on white paper or tracing cloth. Cross-section paper used for graphs should not have more than four lines per inch. If finer ruled paper is used, the major division lines should be drawn in with black ink, omitting the finer divisions. In the latter case, only blue-lined paper can be accepted. Photographs must be very distinct, and must be printed on glossy white paper. Blueprinted illustrations of any kind cannot be used. All lettering should be $^3/_{16}$ in. high for an 8 x 10 in. figure. Legends for figures should be tabulated on a separate sheet, not lettered on the illustrations.

Mathematics—Fractions should be indicated by a slanting line. Use standard symbols. Decimals not preceded by whole numbers should be preceded by zero, as 0.016. Equations may be written in ink with subscript numbers, radicals, etc., in the desired proportion.

Abbreviations—Write a.c. and d.c., (a-c and d-c as adjectives), kc, μf, μμf, e.m.f., mh, μh, henries abscissas, antennas. Refer to figures as Fig. 1, Figs. 3 and 4, and to equations as (5). Number equations on the right in parentheses.

Summary—The summary should contain a statement of major conclusions reached, since summaries in many cases constitute the only source of information used in compiling scientific reference indexes. Abstracts printed in other journals, especially foreign, in most cases consist of summaries from published papers. The summary should explain as adequately as possible the major conclusions to a non-specialist in the subject. The summary should contain from 100 to 300 words, depending on the length of the paper.

Publication of Paper

Disposition—All manuscripts should be addressed to the Institute of Radio Engineers, 33 West 39th Street, New York City. They will be examined by the Committee on Meetings and Papers and by the Editor. Authors are advised as promptly as possible of the action taken, usually within one month.

Proofs—Galley proof is sent to the author. Only necessary corrections in typography should be made. *No new material is to be added.* Corrected proofs should be returned *promptly* to the Institute of Radio Engineers, 33 West 39th Street, New York City.

Reprints—With the galley proof a reprint order form is sent to the author. Orders for reprints must be forwarded promptly as type is not held after publication.

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John V. L. Hogan

Member, Board of Direction, 1929

John V. L. Hogan was born in Philadelphia, Pa., February 14, 1890. He attended Sheffield Scientific School, taking an electrical engineering course. In 1906–1907 he was a laboratory assistant to Lee de Forest in experimental radio telephone work. In 1908–1909 Mr. Hogan was engaged in experimental work at Sheffield Scientific School Graduate Research Laboratory. He joined the staff of the National Electric Signalling Company as telegraph engineer in December, 1909, where he was associated with Professor R. A. Fessenden in a number of research projects. Early in 1912 Mr. Hogan supervised the erection of the Bush Terminal Station in New York City. He had charge of the test operations between the Navy's first high power station at Arlington, Va. and the *USS Salem* in 1913.

Mr. Hogan was appointed chief research engineer of the National Electric Signalling Company in 1914. In 1917 the company's name was changed to International Signalling Company, and Mr. Hogan was made commercial manager. In 1918 he was made manager of the International Radio Telegraph Company. Since 1921 he has been engaged in consulting radio engineering in New York City.

Mr. Hogan has been a prolific contributor to the literature of the art.

His connection with the Institute dates back to his cooperation in the consolidation of the Institute of Wireless Telegraphy and the Wireless Institute, in 1912. He has been a member of the Board of Direction of the Institute continuously since 1913. He served as vice president of the Institute during the period 1916–1919 and was elected president in 1920. Mr. Hogan has been chairman of a number of the Committees of the Institute in the past, including Membership, Publicity, and Standardization. He was elected to full membership in the Institute in 1912, and was transferred to the Fellow grade in 1915.

INSTITUTE NEWS AND RADIO NOTES

1930 Year Book

The membership catalog in the 1930 Year Book will be dated as of December 15, 1929 and will contain the names and addresses of all paid members of the Institute on that date.

If any change has been made in either the business title, business address, or mailing address of a member since the 1929 Year Book was issued, the Institute office should be notified thereof on the form found on page XXXV of the advertising section in this issue, *unless such notification has already been given.*

In the absence of other information the 1930 membership catalog will be identical with that of 1929.

October Meeting of Board of Direction

At the meeting of the Board of Direction of the Institute held at 4 P.M. on October 2nd the following officers and members were present: A. Hoyt Taylor, president; Melville Eastham, treasurer; John M. Clayton, secretary; Alfred N. Goldsmith and Ralph Bown, junior past presidents; W. G. Cady, L. M. Hull, C. M. Jansky, Jr., R. H. Manson, R. H. Marriott, and L. E. Whittemore.

The following were transferred or elected to higher grades of membership in the Institute: transferred to the grade of Fellow: E. E. Bucher, Wm. Dubilier, H. T. Friis, J. W. Horton, D. G. Little, H. B. Richmond, and W. Van B. Roberts; transferred to the grade of Member: E. C. Ballentine, H. W. Baukat, L. S. Brach, Edmond Bruce, C. E. Butterfield, F. E. Canavaciol, R. C. Colwell, P. McK. Deeley, Francis R. Ehle, W. L. Everitt, J. F. Farrington, G. Robert Geit, W. N. Goodwin, Jr., C. W. Hansell, M. P. Hanson, H. D. Hayes, H. H. Henline, C. C. Heselton, W. S. Hogg, Jr., A. A. Howard, I. J. Kaar, G. W. Kenrick, H. M. Lane, C. R. Leutz, H. S. Osborne, and F. X. Rettenmeyer; elected to the grade of Member: E. W. Butler, C. F. Fielding, H. A. Frederick, C. F. Holden, T. H. Kinman, and C. C. Langevin.

One hundred and eighteen Associate members and sixteen Junior members were elected.

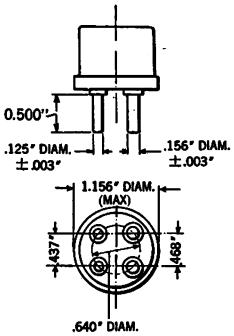
Standard Vacuum-Tube Bases

The Institute of Radio Engineers and the American Institute of Electrical Engineers are joint sponsors of the Sectional Committee on Radio of the American Standards Association. The technical committee on vacuum tubes, C. B. Jolliffe, chairman, has formulated standard specifications for vacuum-tube bases which were formally approved by the American Standards Association on July 26, 1929, as a tentative American standard.

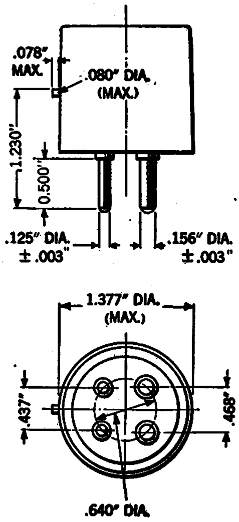
These specifications are as follows:

1. Dimensions

The important dimensions of the standard four-pin vacuum-tube base shall be as shown in the following drawings.


0.500"
.125" DIAM.
±.003"
.156" DIAM.
±.003"
1.156" DIAM.
(MAX)
.437"
.468"
.640" DIAM.

(a) Small Type


.078"
MAX.
.080" DIA.
(MAX.)
1.230"
0.500"
.125" DIA.
±.003"
.156" DIA.
±.003"
1.377" DIA.
(MAX.)
.437"
.468"
.640" DIA.

(b) Large Type

2. Connections

In a triode it shall be standard to connect the tube elements to the four-pin base as shown in the following drawing.

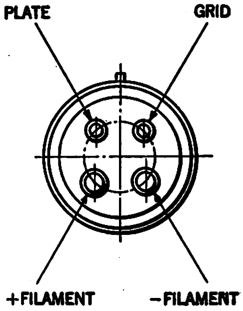


PLATE
GRID
+FILAMENT
−FILAMENT

Single page reprints of these specifications can be obtained from the Secretary of the Institute of Radio Engineers at ten cents a copy.

Institute Meetings

NEW YORK MEETING

At the regular New York meeting of the Institute, held on October 2nd in the Engineering Societies Building, 33 West 39th Street, two papers on aircraft radio were presented.

The first, "Radio for the Air Transport Operator," by Lester D. Seymour, general manager of National Air Transport, of Chicago, was presented by M. M. Eells, communication engineer of that organization. The second paper was presented by the author, H. J. Walls, airways radio engineer, Department of Commerce, Bureau of Lighthouses, Washington. The title was, "The Civil Airways and their Radio Facilities."

Both papers will be published in an early forthcoming issue of the PROCEEDINGS.

A. Hoyt Taylor, president of the Institute, presided at the meeting which was attended by two hundred and fifty members and guests.

ADDRESSES BEFORE SECTIONS BY H. E. HALLBORG

During the month of September, H. E. Hallborg presented a paper, "The Radio Plant of RCA Communications, Inc.," before five sections of the Institute. The meetings were arranged as follows:

Buffalo-Niagara section at the University of Buffalo, September 17th. L. Grant Hector, chairman of the section, presided. Seventeen members were present.

Chicago section at the Electric Club, September 20th. B. B. Minnium, vice chairman of the section, presided. Seventy members of the Institute attended the meeting. The paper was discussed by Messrs. Morse, Minnium, Miller, Oxner, Cole, and Packman.

Cleveland section at the Case School of Applied Science, September 18th. Bruce W. David, chairman of the section, presided. Thirty-seven members of the Institute were present.

Detroit section at Leland-Detroit Hotel, September 19th. A. B. Buchanan, chairman of the section, presided. Fifty members of the Institute were present. A lengthy discussion of the paper followed.

Toronto section in the Electrical Building, University of Toronto, September 16th. V. G. Smith, chairman of the section, presided. The attendance was sixty. The paper was discussed by Messrs. Pipe, Kinnear, Lowry, Dalton, Thompson, Choat, and Richardson.

CINCINNATI SECTION

A meeting of the newly organized Cincinnati section was held on September 16th in the Chamber of Commerce Building, Cincinnati.

Officers of the section were elected as follows: Ralph H. Langley, chairman; W. C. Osterbrock, vice chairman; and W. W. Boes, secretary-treasurer. Mr. Langley announced the appointment of the following committees: Meetings and Papers, Publicity, and Membership.

Two papers were presented at the meeting. The first by Ralph P. Glover was entitled "Notes on Day to Day Variations in Sensitivity of Broadcast Receivers." The second paper by E. T. Flewelling was on "Short-Wave Transmission and Reception." Messrs. Langley, Loftis, and Israel discussed these papers.

Forty-one members and guests attended the meeting.

LOS ANGELES SECTION

The first fall meeting of the Los Angeles section was held on September 16th in the offices of The Southern California Telephone Company, 740 S. Olive Street. Thomas F. McDonough, chairman of the section, presided.

Through the courtesy of Mr. Stannard, division manager of The Southern California Telephone Company, C. E. Lilly, engineer in charge of the telephoto division, gave a lecture covering the operation of the telephoto apparatus. Following Mr. Lilly's talk there was a demonstration of picture transmission.

Due to the unusual interest in this program it was necessary to divide the meeting into groups in order that all could witness the demonstration and hear the lecture.

One hundred and thirty-one members and guests attended the meeting.

PITTSBURGH SECTION

On September 17th a meeting of the Pittsburgh section was held in the Duquesne Light Building, Pittsburgh. L. A. Terven, chairman of the section, presided. Saul Levine, of the Pittsburgh office of the General Electric Company, presented a paper, "Vacuum Tubes," illustrated with lantern slides.

Messrs. J. G. Allen, A. Mag, S. I. Davis, J. P. Bucher, R. L. Davis, A. P. Sumergreen, L. A. Terven, S. Skrypzak, R. D. Wyckoff, and C. B. Upp participated in the discussion which followed.

Thirty-six members of the section attended the meeting.

The paper was prepared by the Research Laboratory of the General Electric Company at Schenectady. It was non-technical in nature and dealt with the fundamentals of vacuum-tube construction and tube applications.

ROCHESTER SECTION

The members of the Rochester section have elected officers for the 1929–1930 season as follows: chairman, Earl C. Karker; vice chairman, Harry C. Gordon; secretary-treasurer (temporary), George Dodson; Board of Directors, Ray H. Manson, Virgil Graham, H. J. Klumb, I. G. Maloff, A. L. Schoen, and A. E. Soderholm.

SAN FRANCISCO SECTION

On September 18th a meeting of the San Francisco section was held in the Bellevue Hotel, 505 Geary Street, San Francisco. Donald K. Lippincott, chairman of the section, presided.

W. F. Frederick presented a paper, "The Screen-Grid Vacuum Tube." The paper described some of the characteristics of the screen-grid tube and pointed out certain advantages and disadvantages of using this tube as a radio-frequency amplifier at broadcast frequencies. A circuit for testing the screen-grid tube was described.

Twenty-five members of the section attended the dinner preceding the meeting, at which forty-three were present.

SEATTLE SECTION

The September 21st meeting of the Seattle section comprised a trip to Keyport, Washington, to the U. S. Naval Transmitting Station. Five of the Naval transmitters, the largest of which is a 30-kw tube set, were inspected. C. E. Williams, naval engineer and a member of the Seattle section, described in detail this transmitter and four others.

The Seattle section chartered the steamer *Atlanta* to make the trip. Thirty-eight members of the section participated in this tour.

WASHINGTON SECTION

At the meeting of the Washington section held on September 12th in the Continental Hotel, Washington, D. C., K. S. Weaver, of the Westinghouse Lamp Company, presented a paper, "Production Testing of Vacuum Tubes."

The paper is summarized as follows: In order to insure the quality of the product of a factory turning out thousands of tubes of a kind every day, a well-organized system of mechanical and electrical inspection is required. Beside the factory inspection, which is made on all of the product, various control inspections, both regular and special, are used to insure the accuracy of the original inspection.

The ability of a tube to stand the rough handling incident to shipping is of prime importance in design. Information on the strength of tubes is obtained by the use of standard drop tests, the change in characteristics due to the dropping being used as a measure of the strength of a tube.

Beside the above outline of test procedure, the paper deals briefly with the relation between the various characteristics and their significance in set operation.

The elementary circuit arrangements used for the measurement of the most important control characteristics are described, and some of the reasons for the use of different circuit arrangements for different tubes are discussed.

Messrs. Davis, Carroll, Guthrie, Wheeler, Case, Brady, Smith, Amick, Wiseman, and Stevens participated in the discussion of the paper.

Vice Chairman Thomas McL. Davis presided at the meeting.

Following the paper and its discussion, motion pictures by the Western Electric Company on the audion and the telephone repeater were shown.

The next meeting of the section will be held on October 10th at the Continental Hotel.

Committee Work

SECTIONS MANUAL

The Board of Direction of the Institute at its October 2nd meeting adopted and approved the "Manual for Section Organization and Operation," a document originally prepared by the 1927 Committee on Sections and revised by the 1928 and 1929 Committees.

The present edition contains much information of assistance to members interested in the formation of a section of the Institute. It also includes the approved form of Constitution for Sections.

Members interested in the possibility of a section in their vicinity, where no existing section now functions, may secure a copy of the manual by addressing the secretary of the Institute.

COMMITTEE ON ADMISSIONS

The regular monthly meeting of the Committee on Admissions was held on October 2nd in the office of the Institute at 9:30 A.M. Members present were as follows: J. S. Smith, acting chairman; A. Hoyt Taylor, C. M. Jansky, Jr., and A. F. Van Dyck.

The Committee considered twenty applications for transfer or election to the higher grades of membership in the Institute, approving nine.

COMMITTEE ON STANDARDIZATION

J. H. Dellinger, chairman of the Committee on Standardization, called a meeting of the four subcommittee chairmen on September 5, 1929, in the Institute office in New York City. Those present were: J. H. Dellinger, Haraden Pratt, chairman of subcommittee No. 2 on transmitters and antennas; H. A. Frederick, chairman, subcommittee No. 4 on electro-acoustic devices; L. E. Whittemore, acting secretary, Sectional Committee on Radio, American Standards Association; and H. P. Westman, assistant secretary of the Institute.

The meeting was held to coordinate the activities of the subcommittees, to define the scope of their respective fields and their relation to the standardization projects of the manufacturers' groups, and to institute the work of the 1929 committee and subcommittees.

H. P. Westman was appointed secretary of the Committee on Standardization.

COMMITTEE ON MEMBERSHIP

A meeting of the Committee on Membership was held at 5:30 P.M. in the Institute office on October 2nd. Those present were: I. S. Coggeshall, chairman; A. F. Murray, F. R. Brick, J. E. Smith, and H. C. Gawler.

The committee is assisting the Committee on Admissions in recommending the names of possible candidates for transfer or election to higher grades of membership in the Institute, having approved some one hundred Associate members for transfer to the full Member grade.

The 1928 Committee on Membership, through its suggestions of eligible members, was mainly responsible for the recent transfer of a large number of Associate members to the Member and Fellow grades.

Personal Mention

J. D. Durkee, formerly officer in charge, Radio Central, Navy Department, Washington, D. C., has recently become chief communications engineer of Universal Wireless Communication Company, Inc., with headquarters at Chicago.

J. G. Mullen has been appointed assistant general manager of the Brunswick-Balke-Collender Company of Canada, Ltd., with headquarters in Toronto. Mr. Mullen was formerly located in Baltimore, Md., with the same company.

L. A. Dye, until recently instructor in mathematics at the University of Rochester, has assumed the position of instructor in mathematics in Cornell University.

Sidney Fishberg, formerly manager of the vitreous enamel resistor department of the Electro-Motive Engineering Corporation of New York City, has joined the staff of Micamold Radio Corporation of Brooklyn as supervisor of the synthetic resistor department.

William H. Gerns has resigned from the Brandes Corporation where he was employed as design engineer to become associated with Wired Radio, Inc., of Ampere, New Jersey, as a designer.

R. P. Glover, a former student of the University of Cincinnati, has joined the radio engineering department of the Crosley Radio Corporation at Cincinnati.

Lloyd T. Goldsmith, until recently research assistant, Massachusetts Institute of Technology at the M. I. T. Experimental Station, Round Hill, Mass., is now associated with the Paramount Famous Lasky Corporation in the sound recording department.

William B. Gould is now with the Western Electric Company as equipment engineer. He was formerly radio engineer at Station WTAG.

L. C. Herndon, formerly U. S. Radio Inspector at Baltimore, has been promoted to the office of Supervisor of Radio.

F. L. Hopper is now an engineer in the recording engineering department of Electrical Research Products, Inc., at Los Angeles. Mr. Hopper was formerly engineer with the Pacific Telephone and Telegraph Company at Los Angeles.

A. H. Hotopp, until recently in the department of development and research in the American Telephone and Telegraph Company, is now associated with Kolster Radio Corporation-Brandes Laboratories, Inc., of Newark, New Jersey, as transmission development and research engineer.

Joseph C. Hromada, formerly a student at Armour Institute of Technology, has been appointed junior radio engineer of the Airways Radio Division of the Department of Commerce at Washington.

Harold R. Hunkins has joined the engineering staff of the International Communications Laboratories of New York City. Mr. Hunkins was formerly a student at Harvard Engineering School of Cambridge, Mass.

George W. Kelley, Jr., has been transferred to the design section, Navy Department, Bureau of Engineering, Washington, D. C. from the Norfolk Navy Yard at Portsmouth, Va.

L. E. West, until recently in the engineering department of the

Grigsby-Grunow Company of Chicago, is now engineer with E. T. Cunningham, Inc., of New York City.

Orville M. Dunning, formerly engineer with Acoustic Products Company of New York City, has recently become chief engineer of Sonora Phonograph Company of Buffalo.

C. K. Jen, until recently with the Moore School of Electrical Engineering, University of Pennsylvania, has become affiliated with the electrical engineering department of Rensselaer Polytechnic Institute of Troy, N. Y.

W. G. McConnell has become engineer with the Southern Radio Corporation at Station WMU, Linden, New Jersey.

E. A. Michelman has joined the engineering and test department of Radio-Victor Corporation, of Camden, New Jersey. He was formerly with the Hammond Radio Research Laboratories, of Gloucester, Mass.

Harold R. Miller recently resigned from the Navy Yard, Boston, Mass., to join the radio engineering department of the Westinghouse Electric and Manufacturing Company at their Chicopee Falls, Mass. plant.

Eugene A. Moore, formerly in the engineering department of Steinite Laboratories at Atchison, Kansas, has joined the engineering staff of the Aircraft Corporation, Atchison, Kansas.

C. P. Sweeny has joined the engineering staff of Bell Telephone Laboratories, New York City.

Arthur F. Van Dyck is now manager of the engineering and test department, Radio-Victor Corporation of America at Camden, New Jersey. Mr. Van Dyck was formerly manager of the technical and test department of R. C. A. in New York City. The latter department has been moved to Camden to constitute the new engineering department.

Earl Arnett has recently assumed the position of chief installation engineer in the General Sound Equipment Company, of Toronto, Ontario.

K. Charlton Black has resigned an instructorship at Harvard University to take the position of physicist with the Boonton Research Corporation at Boonton, N. J.

A. B. Chamberlain has resigned from his position as general manager of broadcasting station WHAM to accept a position with the Buffalo Broadcasting Corporation as chief engineer. Mr. Chamberlain was, at the time of his resignation, chairman of the Rochester section of the Institute.

Edwin B. Dallin, until recently in the sound division of the Naval Research Laboratory at Bellevue, D. C., is now associated with the

Wireless Specialty Apparatus Company, of Boston, Mass., engaged in development and design of radio direction-finders and radio compasses.

A. N. Marzulli, formerly teaching fellow in electrical engineering at Valparaiso University, is now radio engeineer with Temple Corporation of Chicago.

W. A. Murray, formerly assistant professor of engineering in Montana State College, is now assistant professor of electrical engineering at Michigan State College.

Alden C. Packard, graduate student of Harvard Engineering School, has become associated with Heintz and Kaufman of San Francisco as radio engineer.

F. L. Shaw, recently radio editor of the *Providence Journal*, is now associated with United Reproducers' Corporation at Dayton, Ohio, as publicity director.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 11 *November, 1929*

GEOGRAPHICAL LOCATION OF MEMBERS ELECTED OCTOBER 2, 1929

Transferred to the Fellow grade

Massachusetts	Cambridge, 30 State St.	Horton, J. Warren
	Cambridge A, 30 State St.	Richmond, Harold B.
	Chicopee Falls, Westinghouse Elec. and Mfg. Co.	Little, Donald G.
New Jersey	Princeton, 155 Hodge Road	Roberts, Walter van B.
	Rumson	Friis, H. T.
New York	New York City, 411 Fifth Ave.	Bucher, E. E.
	New York City, 10 E. 40th St.	Dubilier, Wm.

Transferred to the Member grade

Dist. of Columbia	Bellevue, Anacostia, Naval Research Laboratory	Hanson, Malcolm P.
Illinois	Chicago, 5746 Drexel Blvd	Deeley, Paul McK.
	Chicago, 2022 the Engineering Bldg	Hayes, H. D.
	Chicago, 6222 Woodlawn Ave.	Heselton, Chas. C.
	Chicago, 306 S. Wabash Ave.	Howard, Austin A.
Maryland	Annapolis, P. G. School, U. S. Naval Academy	Giet, G. Robert
Massachusetts	Belmont, 350 Lake St.	Lane, Henry M.
	Tufts College	Kenrick, Gleason W.
New Jersey	Newark, 127 Sussex Ave.	Brach, Leon S.
	Newark, 30 Scheerer Ave.	Goodwin, W. N., Jr.
	Red Bank, 27 Buena Place	Bruce, Edmond
	Riverton, 900 Main St.	Ehle, Francis R.
	Westfield, 519 Alden Ave.	Baukat, Henry W.
New York	Brooklyn, 326 Leonard St.	Canaraciol, Frank E.
	New York City, Associated Press, 383 Madison Ave.	Butterfield, C. E.
	New York City, Intl. Com. Lab., 67 Broad St.	Farrington, John F.
	New York City, 33 W. 39th St.	Henline, Henry H.
	New York City, 195 Broadway	Osborne, H. S.
	New York City, c/o Postmaster, *USS Hannibal*	Hogg, W. S., Jr.
	Port Jefferson, 115 Tuthill St.	Hansell, C. W.
	Schenectady, 2024 Euclid Ave.	Ballentine, Edwin C.
	Schenectady, Radio Eng. Dept., General Electric Co.	Kaar, Ira J.
	Woodside, L. I., 41-06 Fiftieth St.	Rettenmeyer, F. X.
Ohio	Columbus, Dept. of Elec. Eng., Ohio State University	Everitt, William Littell
Pennsylvania	Altoona	Lentz, Charles Roland
West Virginia	Morgantown, 332 Demain Ave.	Colwell, Robert Cameron

Elected to the Member grade

California	San Francisco, c/o E. T. Cunningham, 182 2nd St.	Butler, E. W.
	San Francisco, 274 Brannan St.	Langevin, Carl C.
New York	New York City, USS *Wyoming*, c/o Postmaster	Fielding, Charles F.
	New York City, 463 West St.	Frederick, Halsey A.
	New York City, c/o Postmaster (U. S. Naval Mission to Brazil)	Holden, Carl F.
England	Rugby, Bilton Road, "Ladbrooke" House	Kinman, Thos. H.

Elected to the Associate grade

California	Beverly Hills, Fox Film Co.	Larsen, C. W.
	Covina, 165 No. Covina Blvd.	Ross, M. F.
	Culver City, Hotel Washington	Wersen, David T.
	Huntington Park, 2464 Randolph St.	Kirby, Melvin D.
	Los Angeles, 1839 E. Vernon Ave.	Tami, Joseph, Jr.
	Oakland, 7515 Weld St.	Madison, C. E.
	Oakland, 5555 E. Fourteenth St.	Parkhurst, Edgar L.
	Palo Alto, c/o Federal Telegraph Co.	Harrison, Charles I.
	San Diego, 4558 32nd St.	Farnum, Willis H.
	San Francisco, 242A Hartford St.	Arrigoni, Arthur
	Sawtelle, 1446 Saltair Ave.	Spiller, Cecil Charles
Connecticut	West Haven, 12 Ward Place	Phillips, Ed. I.
Dist. of Columbia	Washington, Hqtr's U. S. Marine Corps	Cole, G. C.
	Washington, Naval Research Laboratory	Hentschel, Ernest R.
	Washington, 4107 Ingomar St., N. W.	Norton, Kenneth A.
	Washington, 226 8th St., S. E.	Wilkie, Harry
Illinois	Chicago. 4111 Ravenswood Ave., E.	Beard, J. Gregson
	Chicago, 744 Addison St.	Gay, Paul F.
	Chicago, 5317 Argyle St.	Hansen, Harvey Bennett
	Chicago, 1532 S. Homan Ave.	Levin, Sam
	Chicago, 2005 Prairie Ave.	Lopes, John
	Chicago, 1865 Daily News Plaza	McClintock, W. S., Jr.
	Chicago, 39 N. La Salle St., Room 1117	Scharf, Joachim Barschach
	Chicago, 4536 Magnolia Ave.	Shultise, Q. M.
	Chicago, 3547 Pierce Ave.	Thorsen, Orville T.
	Chicago, 4003 N. Kildare Ave.	Weibler, Carleton T.

Illinois (cont'd)	Great Lakes, U. S. N. Radio Station	Clark, Thomas F.
	Naperville	Stoos, Jos. A.
	Oak Park, 701 S. Kenilworth Ave	Weiser, Carl
	Streator, 210 W. Lincoln Ave	Melody, Bernard J.
Indiana	Valparaiso, P. O. Box 292	Norris, Sam
Iowa	Onslow	Koon, Cecil L.
Louisiana	New Orleans, c/o Tropical Radio Tel. Co., 321 St. Charles St.	Nall, Vance
Maine	Portland, 67 Bradley St.	Ryall, Henry
Maryland	Baltimore, 1626 Warwick Ave.	Cohen, H. A.
Massachusetts	Cambridge, General Radio Co., 30 State St.	Bousquet, Arthur G.
	Dorchester, 65 Monadnock St.	Hardy, Carroll N.
	Mattapan, 55 Goodale Road	Parnes, Henry
	Southbridge, 53 Oliver St.	Yates, Wilfred
Michigan	Detroit, 11391 Marlowe Ave.	Geiger, Arthur H.
	Flint, 830 Paddington Ave.	Lutos, Clifford
	Jackson, 214 N. Pleasant St.	Fortier, Raymond C.
	Jackson, R. R. #1, Harding Road	Lee, Kenneth G.
	L'Anse	Fridgen ,Edward N.
Minnesota	Minneapolis, 3519 24th Ave., So.	Kelly, James, Jr.
Missouri	Jefferson City, Radio Station WOS	Sloan, Fergus M.
	Kansas City, 6500 Paseo	Rippeteau, Chas. Wm.
New Jersey	Ampere, P. O. Box 58	Walz, Richard F.
	Deal, P. O. Box 122	Goodall, William M.
	East Orange, 44 Clifford St.	Pickard Richard W.
	Lawrenceville, Box 132	Schwartz, Lyle H.
	Nutley, 31 Burnett Place	Humphrey, Hartley C.
	West Collingswood, 425 Taylor Ave.	Pettit, Albert R.
	West Orange, 2 Mountain View St.	Jenkins, Robert
New York	Astoria, L. I., 2255 33rd St.	Kunicky, Barney F.
	Brooklyn, 2977 W. 3rd St.	Daniels, Lew
	Brooklyn, 734A Fourth Ave.	Francione, Dominick A., Jr.
	Brooklyn, 180 Driggs Ave.	Mullane, John W.
	Brooklyn, 1872 Douglass St.	Peck, William
	Brooklyn, 689 Lenox Road	Renke, Adolph
	Brooklyn, 1430 Flatbush Ave.	Salzer, Arthur H.
	Brooklyn, 64 Linden St.	Thomas, Edward H.
	Brooklyn, 2076 66th St.	Widmann, Erwin
	Jamesport	Seaman, James Corwin
	Long Island City, 481 Grand Ave.	Holzinger, Theo. E.
	Middletown, 174 W. Main St.	Dempsey, William E.
	Mt. Vernon, 8 Beekman Ave.	Macalpin, William W.
	New York City, Bell Tel. Labs., 463 West St.	Bauer, Brunton
	New York City, Bell Tel. Labs., 463 West St.	Boesche, F. W.
	New York City, 2748 Holland Ave.	Boltson, Jacob A.
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	New York City, c/o E. T. Cunningham, Inc., 370 7th Ave.	Carroll, Michael J.
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	New York City, 3985 Saxon Ave.	Gihring, Herman E.
	New York City, 261 5th Ave., Room 1800	Gladkov, Cyril A.
	New York City, 1056 Boston Road	Hirsch, Harry
	New York City, Bell Tel. Labs., 463 West St.	Hudack, John Martin
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	New York City, 132 E. 36th St.	McSweeny, Roger
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	Schenectady, General Electric Co., Room 419	Ferris, Warren Robert
	Schenectady, General Electric Co., 1 River Road	Thompson, B. J.
North Carolina	Raleigh, c/o Radio WPTF	Newman, John W.
Ohio	Canton, 1030 Clarendon Ave., S. W.	Ellis, Walter R.
	Cleveland, 2049 Cornell Road	Fowler, J. Randall
	Dayton, 521 Negley Place	De Weese, Herbert William
Oklahoma	Asher, c/o Prairie Pipe Line Co.	Lewis, John B.
	Oklahoma City, 103 W. 13th St.	Miller, Wayne
	Oklahoma City, 1633 W. 11th St.	Stokes, Ray
Oregon	Salem, 555 Belmont St.	Churchill, H. B.
Pennsylvania	Lansdowne, 260 Green Ave.	Warren, S. Reid, Jr.
	Philadelphia, c/o Howson & Howson, 123 So. Broad St.	Cerstvik, Stephen
	State College, Acacia Fraternity	Long, Marvin
Rhode Island	Providence, Brown University	Andrews, Howard L.
	Wakefield	Taylor, Alfred R.
Texas	Tyler, Tyler Commercial College	Lowrey, Byron G.

Canada	Kingston, 164 Queen St.	Tanner, Chas. J.
	Montreal, 835 Ave. Laurier est	Tremblay, Jas.
	Toronto, Ont., 64 Hayden St.	Campbell, Henry Lawson
Channel Islands	Guernsey, King's Road, Millbrook	Manning, William Montagu
England	Dorchester, Dorset, Marconi's Beam Transmitting Station	Clarke, Douglas F.
	Plymouth, 49 Greenbank Ave.	Curd, David A.
	Sunderland, Southwick, 12 Dryden St.	Joice, W. S.
	Wisbech, Cambs, Gorefield, "Palestine"	Holmes, Cyril T.
India	Bangalore, Hebbal Post. Indian Institute of Science, Dept. of Elec. Technology	Doraswamy, M. N.
New Zealand	Auckland, Mt. Albert, 9 Veronica Ave.	White, Russell G.
Philippine Islands	Manila, Philippine School of Arts and Trades	del Rosario, Manue S.
South Africa	Klipheuvel, C. P., Wireless Station	Nutt, A.
	Salisbury, S. Rhodesia, Automatic Exchange, Box 391, G. P. O.	Jephcott, Ernest L.
South Australia	St. Peters, 42 Nelson St.	Linnstt, Douglas N.

Elected to the Junior grade

Colorado	Durango, P. O. Box 594	Dieckman, Wm.
Illinois	Carbondale, 800 W. Pecan St.	Chapman, David E.
	Chicago, 4509 N. Robey St.	Chadwick, Ray E.
	Chicago, 7404 Bennett Ave.	Nardin, George F., Jr.
	Litchfield, 224 Van Buren St.	Weller, Earl Selwyn
Kansas	Wichita, 605 Laura Ave.	Demuth, G. W.
Massachusetts	Brockton, 17 E. Ashland St.	Sampson, Edward J. F.
	Worcester, Y. M. C. A., Room 319	Anderson, Paul E.
Missouri	St. Louis, 946 Belt Ave.	Martin, J. Douglas, Jr.
New York	New York City, 3164 Grand Concourse	Linde, James E.
Pennsylvania	Philadelphia, 224 E. Sharpnack St.	Bosco, Joseph F.
	Upper Darby, 6994 Ruskin Lane	Slattery, John J.
South Dakota	Sanator	Tilgner, Shelton R.
India	Benares City, 56 Luxa	Ghosh, B. N.

* * *

Proceedings of the Institute of Radio Engineers
Volume 17, Number 11 *November, 1929*

APPLICATIONS FOR MEMBERSHIP

Applications for transfer or election to the various grades of membership have been received from the persons listed below, and have been approved by the Committee on Admissions. Members objecting to transfer or election of any of these applicants should communicate with the Secretary on or before November 30, 1929. These applicants will be considered by the Board of Direction at its December 4th meeting.

For Transfer to the Fellow grade

Ohio	Cincinnati, 5449 Hamilton Ave.	Langley, Ralph H.

For Transfer to the Member grade

California	Hollywood, 1041 N. Formosa Ave.	Hilliard, John K.
New Jersey	East Orange, 102 N. Arlington	Towner, Orrin W.
New York	New York, Western Union Tel. Co., 195 Broadway.	Coggeshall, Ivan S.
Pennsylvania	Wilkinsburg, 435 Avenue D	Beers, G. Lisle

For Election to the Member grade

New York	New York City, Dept. of Commerce, Radio Service, Subtreasury Bldg.	Fish, Arthur S.
Philippine Islands	Manila, Fort Santiago, Radio Cable and Telegraph Office	Rives, Tom C.
Poland	Warsaw, Wspolna 32	Plebanski, Jozef
	Warsaw, Mokotowska 6, Institute Radiotechnique	Sokolcow, Dmitri M.

For Election to the Associate grade

Alabama	Ensley, 3000 Pike Ave.	Sartain, L. B., Jr.
California	Los Angeles, 3686 S. Main St.	Dale, James Albert
	Los Angeles, 315 La Bertha Apts., 960 S. Oxford Ave.	Dickson, H. L.
	Los Angeles, 563 N. Hoover St.	Olmstead, Chas. B.
	Palo Alto, 430 Addison Ave.	Morgan, Nathaniel R.
	Palo Alto, Federal Telegraph Co.	Saunders, Wm. W.
	Palo Alto, 1119 Fulton St.	Thomas, Chas. C.
	Palo Alto, 603 Fulton St.	White, Ray H.
	Palo Alto, c/o Federal Telegraph Co.	Whitwam, Lloyd F.
	San Francisco, 285 Turk St.	Schofield, William D.
	Southgate, 8148 Virginia St.	Ott, Ben A.
Dist. of Columbia	Takoma Park, 218 Cedar St.	Mertie, J. B., Jr.
	Washington, Tower Bldg.	Jardine, W. N.
	Washington, Radio Section, Bureau of Standards	Shankland, Robert S.
Illinois	Chicago, 6900 Constance Ave.	Lahman, Wilford C.
Indiana	Elkhart, 340 W. Crawford St.	May, Frank Darrell
Iowa	McGregor, Box 282	Woods, Talmage DeWitt
Louisiana	New Orleans, 2430 Gen. Taylor St.	Lobe, Henry, Jr.
Maryland	Annapolis, R.F.D. #1, Box #41A	North, J. Harold
Massachusetts	Revere, 36 George St.	Gelardi, Matthew
	Springfield, 2713 Main St.	Wilson, R. A.
Michigan	Detroit, 15412 E. Jefferson Ave.	Krause, Harold F.
	Detroit, 2909 David Stott Bldg.	Rollins, G. K.
	Scottville	Krivitzky, George
Mississippi	Greenwood, 800 Henry St.	Johnson, Vivion A.
New Jersey	Hillside, 121 Clark St.	Short, William P.
	Maywood, 96 Washington Ave.	Kersta, Laurence G.
	Merchantville, Greenluyh Court Apt. C-9	Lamont, J. Cutler
	Newark, Brandes Laboratories	Oman, Nils Johann
	East Orange, 28 Irving St.	Cooper, John W.
	Passaic, 23 Passaic Ave.	Rodgers, George H.
	Newark, Brandes Laboratories, Inc., 200 Mt. Pleasant Ave.	Gordon, Malcolm K., Jr.
New York	Babylon, L. I., 194 E. Main St.	Thompson, Edward Philip
	Brooklyn, 1197 E. 34th St.	Castle, Donald Hewitt
	New York City, Western Union Telegraph Co., 195 Broadway	Arnold, John W.
	New York City, Room 505, 89 Broad St.	Beizer, Harold
	New York City, c/o Postmaster, *USS Galveston*	Daspit, L. Randall
	New York City, Room 1915, 195 Broadway	Franklin, L. W.
	New York City, Bell Telephone Labs., 463 West St.	Hartman, Charles D.
	New York City, 2675 Valentine Ave., Bronx	Lake, Daniel A.
	New York City, 358 Wadsworth Ave.	Venable, Richard Neel

New York (cont'd)	New York City, 1115 College Ave., Bronx	Wettermann, John A.
	Riverhead, L. I., c/o Radio Corporation of America	Smith, Arthur Z.
	Rockville Center, 358 Village Ave	Schabbehar, Edwin A.
	Schenectady, 325 Van Vranken Ave	DeNardo, Federigo
	Schenectady, 205 Seward Pl	Dickinson, Theodore M.
	Schenectady, 509 Craig St	Somers, Brock A.
	Schenectady, 109 Furmon St	Zimmerman, Arthur G.
	Scotia, 136 Glen Ave	Smith, J. P.
Ohio	Cincinnati, 3317 Bishop St., Apt. 210	Baird, Robert E.
	Cincinnati, 103 Inwood Terrace	Nunneker, William H.
	Cleveland, 1280 Norwood Road, N. E.	Worst, J. S.
	Columbus, 400 W. 7th Ave	Crisante, Aldo
	Columbus, Dept. of Elec. Eng., Ohio State University	Rosenfeld, Millard
	Dayton, 16 Bremen St	Friedman, Harry
Pennsylvania	Allentown, 953 Cedar St	Williams, Thomas R.
	Allentown, 204 S. Fulton St	Young, Charles S.
	Emporium, 301 Poplar St	Brophy, Thomas L.
	Harrisburg, 17 Evergreen St	Howe, David B.
	Harrisburg, 603 Benton St	Kilheffer, Harold E.
	Harrisburg, 1008 N. 3rd St	Knerr, G. Russell
	Harrisburg, 344 Hamilton St	McGary, Oliver H.
	Irwin, Maple St	Brook, Alf.
	Philadelphia, 3421 A St	Barton, R. W., Jr.
	Philadelphia, 243 S. Upsal St., Mt. Airy	Blackwood, George C.
	Philadelphia, Navy Yard, Bldg. 67	Holden, Ellis Gray
	Philadelphia, 1813 Green St	McCloskey, James Walter
	Philadelphia, Gladstone Hotel, Apt. 26	Pickett, C. E.
	Pittsburgh, 733 Chislett St	Diamond, Hymen
	Pittsburgh, 1636 Broadway Ave	Gemmell, William H.
	Pittsburgh, 7 Verner Court, N. S.	Murray, Alex H.
	Pittsburgh, 6801 McPherson Blvd	Nichols, Harry J.
	Pittsburgh, 7211 Thomas Blvd	Overholt, Ralph, Jr.
	Pittsburgh, 530 Kelly Ave.—21	Richardz, Folke
	Pittsburgh, 301 South Ave., Wilkinsburg Branch	Seaverson, Oswald I.
	Pittsburgh, 915 Vista St., N. Side	Wittgartner, J. S.
	St. Mary's, Speer Carbon Co.	Abbott, Harold W.
	Swissvale, Union Switch and Signal Co., Research Dept	Bossart, Paul N.
	Turtle Creek, 310 James St	Gunby, O. B.
	Turtle Creek, 160 Brown Ave	Seabert, J. D.
	Turtle Creek, 131 Grant St	Walker, M. C.
	Verona	Wildow, Paul
	Wall	Popsack, Andrew A.
	Wilkinsburg, 8021 Susquehanna St	Aggers, C. V.
	Wilkinsburg, 206 Union St	Ellis, Grenville Brigham
	Wilkinsburg, 1114 South Ave	Gillette, K. G.
	Wilkinsburg, 1643 Maplewood Ave	Grundmann, Gustave L.
	Wilkinsburg, 841 Holland Ave	Karns, Melvin E.
	Wilkinsburg, Westinghouse Club	Upsahl, E.
	Wilkinsburg, 1318 Singer Place	Simpson, Virgil R.
Texas	Waco, 520 Austin Ave	Barnes, Wilmer N.
Washington	Seattle, 904 E. Pike St	Gable, M.
Wisconsin	Milwaukee, 1007 Murray Ave	Paul, Geo. Stewart
	Plymouth, 914 Eastern Ave	Limberg, Raymond A.
Canada	Toronto, Ont., 2373 Bloor St., W., Apt. 3	O'Brien, J. E.
China	Shanghai, Chinese Govt. Elec. Works, 862 Ave. Haig	Tsiang, Cheng-Yen
England	Cardiff, 8 New St., c/o B. W. M. S. Ltd	Yates, Ivan
	Sandenstead, Surrey, 24 Florence Road	Clear, K. Y.
Hawaii	Honolulu, Wailupe, U. S. Naval Radio Station	Hammond, W. Murray
	Honolulu, 80 Merchant St	Hoover, J. F.
Holland	Hilversum, 1 Arnbalaam	Bouman, J. A. J.
Italy	Milano, Corso Sempione 95	Bacchini, Cesare
Nova Scotia	Halifax, c/o Western Union Tel. Co., Cyrus Field	Nelma, Arsene
Scotland	Edinburgh, 6A Royal Circus	Gee, Harry L.
South Africa	S. Rhodesia, Salisbury, Automatic Exchange Box 391	Jephcott, Ernest L.

For Election to the Junior grade

California	Los Angeles, 1136 S. New Hampshire Ave	Bilkie, W.
Georgia	Atlanta, P. O. Box No. 742	Barker, Howard J.
	Atlanta, 65 Roswell Road	Stapp, Jack
Illinois	Chicago, 1758 Sunnyside Ave	Allain, Joseph A.
Massachusetts	Mattapan, 12 Tennis Road	Rothberg, Joseph
Michigan	Detroit, 8959 Clarion	Ludvigsen, Leonard E.
Nebraska	Lincoln, 117 S. 17th St	Koch, J. Wesley
New York	New York City, 1662 Hoe Ave., Bronx	Flaum, Joe
Pennsylvania	Turtle Creek, 839 Maple Ave	Guzik, Joseph
England	Neasden, Middlesex, "Porthmeor," West Way	Floyd, William F.

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PART II

TECHNICAL PAPERS

TECHNICAL ACHIEVEMENTS IN BROADCASTING AND ITS RELATION TO NATIONAL AND INTERNATIONAL SOLIDARITY†

By

Symposium Committee of the Institute of Radio Engineers

Prepared for World Engineering Congress
Tokio, Japan, October, 1929

1. Introduction, by Alfred N. Goldsmith.
2. Radio Broadcasting Transmitters and Related Transmission Phenomena, by Edward L. Nelson.
3. The National Broadcasting Company—A Technical Organization for Broadcasting, by Julius Weinberger.
4. Speech Input Equipment, by D. G. Little.
5. Wire Line Systems for National Broadcasting, by A. B. Clark.
6. Radio Broadcasting Regulation and Legislation, by J. H. Dellinger.

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†The Institute of Radio Engineers was invited by the American Committee of the World Engineering Congress to participate in the technical program of the World Engineering Congress, held in Tokio in October, 1929, and was asked to present a paper on "Technical Achievements in Broadcasting and Its Relation to National and International Solidarity."

The Board of Direction of the Institute accepted this invitation and appointed a Symposium Committee with membership as follows: Alfred N. Goldsmith, chairman; Edward L. Nelson, Julius Weinberger, D. G. Little, A. B. Clark, and J. H. Dellinger.

The papers published in the following pages are the individual contributions from members of the Committee.

C. W. Latimer, who was present as one of the Institute's delegates, presented the papers before the Congress on behalf of the Institute.

Proceedings of the Institute of Radio Engineers
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INTRODUCTION*

By
ALFRED N. GOLDSMITH
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A GROUP of radio engineers, upon invitation of the American Committee of the World Engineering Congress to be held at Tokio in 1929, has arranged to present to the Congress a related series of papers on various technical aspects of radio broadcasting. Each of the contributing authors has chosen a section of the subject in which he has specialized, and has presented in his contribution to this symposium a summary of past achievements, present problems, and some of the probable lines of future development of this division of radio broadcasting. In the aggregate, therefore, the symposium papers constitute a general summary of the more important technical aspects of radio broadcasting, though necessarily in compact form and with consideration only for major questions.

In this general introduction to the collected symposium papers, it has seemed advisable to restrict the discussion to certain matters of general importance to any nation in which broadcasting is now carried on. More particularly it is self-evident that national solidarity and radio broadcasting have some definite and vital relations to each other;—a closer examination of the nature of these relations may therefore constitute a suitable preface to the symposium papers.

The radio instrumentalities and methods to be described in these papers are the component elements of the greatest system of mass communication which has as yet been evolved by mankind. The possible accomplishments of this method of mass communication are necessarily limited by the performance of the apparatus and systems in question; and accordingly the future increase in the influence of radio broadcasting will measurably depend upon the technical progress which can be expected. It is therefore also appropriate in this introduction to consider the relationship between the quality of national broadcasting service and the engineering devices and methods which are available for such service.

1. Influence of Transmitter Power on National Broadcasting Service

It has apparently been believed by some that an indefinite increase in the power of transmitting stations would in each case result in a corresponding improvement in broadcasting service. When only the low powers heretofore used for broadcasting (0.5 to 5 kw) were under consideration, such a belief was comprehensible. However, in order to reach each member of a great audience with clear and acceptably powerful signals, it has now become evident that powers from several tens of kilowatts to hundreds of kilowatts are required. Nevertheless, a limitation of economically feasible and technically desirable power results from the existence of fading phenomena in radio transmission on middle frequencies which, first intruding at distances of approximately 75 to 150 miles (120 to 240 km) from the transmitter, render any increase of power of limited value for improving service in the "fading zone." Whenever fading is so rapid in such regions as to cause quality distortion, an increase of transmitter power can contribute very little to the situation. On the other hand for regions inside the fading area, or for regions considerably outside the rapid fading area (say at distances of from 300 to 400 miles, or 500 to 700 km), increase of power does in general result in an improvement in service, and particularly for the rural listeners. The criterion governing the amount of power to be used is that the transmitter shall lay down a signal which will over-ride interference, either man-made or natural, at the inner boundary of the fading zone without, however, producing excessive local interference near the transmitter (assuming the transmitter to be appropriately located in a relatively sparsely populated neighborhood). As previously stated, it would appear that powers as high as hundreds of kilowatts will be required for broadcasting service capable of meeting such standards. This result should not be entirely unexpected when it is realized that approximately equal transmitter powers are required for reliable telegraph communication over any considerable distance, even with skilled receiving personnel equipped with elaborate receiving apparatus and antennas.

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* Dewey decimal classification: R550. Original manuscript received by the Institute, January 4, 1929.

2. Influence of the Assignment of Definite Frequency Bands to Individual Transmitting Stations

The healthy national (and international) expansion of broadcasting is dependent upon an orderly arrangement of broadcasting stations, not only in each country, but in neighboring countries as well. If stations are not assigned to definite frequency bands or channels, or if these channels are haphazard in their arrangement, numerous cases of local interference will result, and radio reception in large areas may well become repellent rather than attractive to the listeners.

For this reason the study of available frequency channels for broadcasting, and their assignment to individual stations in a systematic and carefully planned fashion, are among the major factors in rendering radio broadcasting useful to a nation.

3. Influence of the Network Syndication of Programs on National Broadcasting

It is clear that program-producing facilities will in general be unevenly distributed throughout a nation. Frequently most of the artistic and literary ability suitable for entertainment programs of high quality is concentrated in a relatively small number of large cities. This results from economic and group-concentration causes. If the programs originating in such centers of available program material and showmanship are broadcast only locally, it is obvious that a large portion of the country of average size will fail to receive the major benefit of broadcasting. Without detracting from the value of purely local programs from certain stations, it is clear that network syndication of programs (for example, by wire-line distribution or its equivalent to a group of outlet stations all radiating these programs) is a definite national necessity. It enables the coordinated distribution on a large scale of programs which would otherwise be economically unjustified or physically unavailable except in a few portions of the country.

4. Influence of Increase in Fidelity of Tonal Reproduction in Receivers

One of the limitations of radio broadcasting is found in the quality of reproduction of the receiving set and the loud speaker which may be furnished to the individual listener. Inasmuch as the intelligibility of speech and the quality and naturalness of music are determined by the fidelity of reproduction, the influence and importance of broadcasting to a nation is directly related to the fidelity of the average radio apparatus employed by its citizens. The conclusion may be drawn that the use of obsolescent or obsolete receiving sets by any considerable portion of the population of a country will greatly reduce the influence of radio broadcasting and will warrant strenuous efforts to improve the quality of such inferior receiving equipment.

5. General Technical Aspects of Transmitter Service Area, of Receiver Selectivity, of Network Distribution, and of Receiver Fidelity

Inasmuch as the national influence of broadcasting is dependent upon the extent to which its message successfully reaches great masses of people, and since this extent may be evaluated more or less quantitatively, it is desirable to consider more specifically each of the preceding aspects of radio broadcasting.

The service rendered by a broadcasting station is satisfactory to the population only within a region defined as the "service area." It is difficult to specify exactly the limits of the service area. Rural listeners have been known to be content, or at least not openly rebellious, when the radio service delivered to them corresponds to field strength of received signals of as little as 0.01 mv per meter. It is quite certain, even in the rural districts where man-made electrical disturbances are at a minimum, that such field strengths cannot long continue to satisfy the listener.

Up to a few years ago, it was believed that listeners in urban districts should be well satisfied with field strengths of the order of 10 mv per meter. This value was then thought to be adequate to overcome local man-made electrical disturbances and natural summer atmospheric disturbances. Urban listeners have shown an increasing tendency to regard such signal field strengths as inadequate for satisfactory service; and in some districts appear to regard strengths from 30 to 100 mv per meter as not only reasonable but highly desirable.

The field strength at which interference became objectionable in receiving sets of limited selectivity, such as were used a few years ago, was approximately 20 mv per meter. There is no question that field strengths from 100 to as much as 1,000 mv per meter are now at least tolerated in limited districts in the neighborhood of transmitting stations, particularly if such stations are giving excellent service to large areas.

In summary, it would appear that a full utilization of radio broadcasting by any nation involves acceptance at some reasonable date by each nation of the principle that the field strengths required for rural service are between 1 and 10 mv per meter, and for suburban and urban service from 20 to 100 mv per meter, and that objectionable interference begins at field strength values well in excess of 100 mv per meter. It is not to be expected that the radio art will accept such standards immediately, but they are obviously along the trend of present development.

In the second place, the assignment of definite channels to transmitting stations specifies, in turn, the requisite selectivity. Receivers should be sufficiently selective to enable the separation of all local stations in a given district (unless the listener is located extremely near to one or more of the stations), but should not attenuate the sidebands of individual transmissions to such an extent as to injure noticeably program quality.

In considering the best methods of syndicating programs over wire networks, the existence of the fading zone of each station of adequate power will determine in some measure the proper location of the adjacent outlet stations. Presumably each pair of outlet stations should be so located that the overlap of the service areas of such adjacent stations does not constitute a high percentage of the area served by either station. In general, this will require a separation for stations having a power of the order of 100 kw of something between 100 and 200 miles (160 to 320 km) although local conditions (nature of intervening territory, ground and building absorption, distribution of population, and the like) will require consideration in determining the most effective location of the outlet stations.

In considering the fourth point of fidelity of reproduction by the receiver, it should be remembered that the obvious aim of broadcasting is to create an approximation to the "illusion of reality" in the home of the listeners. That is, the speech or music reproduction should be such that, so far as possible, it gives the impression of the original studio rendition. As yet radio engineering has not enabled more than a fair approximation of this standard to be commercially available. The expansion of national broadcasting involves a continuous raising of standards in this direction among the listeners at large, many of whom are in the unfortunate position of using equipment incapable of giving any fair idea of the quality of programs transmitted by modern broadcasting stations and received on high-grade receivers.

6. Influence of National Conditions on the Success of National Broadcasting

Inasmuch as it is desired to reach every listener in a country consistently with adequate signal strength, national broadcasting can be best carried out under the following circumstances:

(a) On such transmitting frequencies as will minimize fading and static. The lowest frequencies in the generally used broadcasting band from 550 to 1500 kc are, therefore, suitable for high-power transmission on cleared channels, particularly when absorption in urban areas (which increases rapidly with the transmission frequency) is taken into account.

(b) Over a terrain of such character that no highly absorbing ground conditions exist and that no extensive mountain ranges intervene. Certain types of territory have a marked attenuation for radio waves and render it extremely difficult or impossible to provide a uniform signal distribution to the listeners. At best, it is a serious problem to supply a reasonably constant signal strength to all listeners at approximately the same distance from the transmitter.

(c) In countries of comparatively small extent. If a country having an area of several million square miles must be supplied with broadcasting service, and if portions of this country are sparsely settled, or are of mountainous or desert character, the problem of distribution becomes highly complicated and involves careful planning to give a reasonably satisfactory service on an equitable basis to most listeners.

(d) If the entire country lies within a single time zone. When on the other hand, there is for example a three-hour time difference between the eastern and western sections of any country, there will be but one hour available between 8 o'clock and midnight each evening (which is the most desirable broadcasting period) when all listeners will be able to receive a given program. Obviously, economic considerations make it difficult to distribute sponsored programs (of advertising character) over areas where considerable time differences exist. Time differences reduce the maximum population which can be simultaneously reached by a national or international program, though to a less extent on holidays.

(e) In countries having nearly uniform population density. If large areas in a country are sparsely populated, it will be difficult to serve such areas economically since most of the power transmitted by radio to such regions will be lost. On the other hand, if certain regions include extremely densely populated areas, in general the steel construction in such areas will result in marked attenuation of the signal waves and great difficulty in giving uniform service to the listeners in such regions. Furthermore, such densely populated areas, in general, have a high noise level due to man-made electrical interference. The ideal location of a transmitting station is in a region with scant population within five or ten miles outside of which lies an area of approximately uniformly populated territory. Of course, such conditions do not exist in general; and a compromise location for the transmitting station must be sought in each instance.

7. Influence of Various Systems of National Program Distribution

Assuming as an ideal that it is desired to reach every broadcast listener with reliable program service, there have been proposed a number of possible methods. The utilization in most countries of a single large station to cover the entire country is clearly not feasible because of the limitation of service area resulting from fading phenomena. A great multiplicity of small stations, some of which at least are operated on exactly the same radio frequency (synchronized stations) has also been proposed. The economic limitations imposed by wire-line network costs make such a plan apparently unsuitable for countries of any considerable size.

The present probable trend, for each national network, is toward comparatively high-powered stations (of from fifty to several hundred of kilowatts in the antenna), with perhaps a limited number of smaller synchronized stations operating on the same frequency as the nearest high-powered station and intended primarily to "fill-in" certain highly attenuating districts where the absorption of the radio waves is so high as to preclude reasonably uniform service for any reasonably distant station utilizing an economic amount of transmitter power. Depending upon terrain conditions, areas of from 10,000 to 50,000 square miles (25,000 to 130,000 square km) can be reasonably well served by such a single station.

In order to provide program diversity, two or more groups of such stations have been established or are planned in certain countries, enabling each coordinated group to send out a different network program. This system of multiple network operation of groups of high-powered stations (with perhaps a limited number of synchronized smaller stations to supply service in regions of unusually high attenuation) represents the latest available practice in national program distribution.

8. National Influence of Various Types of Programs

An examination of radio programs at once discloses a line of demarcation between purely local and definitely national programs. Local programs may be broadly defined as those appealing to a limited number of listeners, in a certain locality. They have a personal quality which makes them attractive to such persons, but of comparatively little interest outside of the limited district in question. Clearly such programs are unsuited for national network syndication.

In contra-distinction, there are certain types of entertainment and educational programs of relatively impersonal nature which, nevertheless, appeal broadly to the entire population of any fairly homogeneous country. Such programs generally originate in the centers of culture or entertainment of the country in question and, if carefully planned, are suitable for national distribution.

It is clear that the latter type of program is of importance in the promotion of national solidarity whereas the local program, despite its personal appeal, readily tends toward a purely provincial character. How far each form of program should be encouraged involves a delicate decision as to the relative merits of programs which stimulate national consciousness and those which, while undoubtedly picturesque in many instances, nevertheless, are essentially local or provincial in character. Perhaps a suitable compromise would involve the occasional introduction of a local type of program over the national networks, whereby the entire population of a country may become aware of the peculiar and frequently likable characteristics of a section of the population.

9. General Classification of Programs and Their National Aspects

There are many possible systems for the classification of programs, most of which are more philosophical or ultra-professional than practical. Programs are also rendered comparatively difficult of classification because the boundary lines between various types of programs are necessarily nebulous and individual programs may include a number of diverse constituent elements.

Entertainment programs appear to be most popular and generally comprise musical material, accounts of sporting events, radio dramas or comedies, and the like. When carefully planned, they are a powerful force for the cultivation of a national musical and artistic taste and, by carrying these various types of entertainment even into remote parts of a country, they are bound to reduce any gap in culture existing between different sections of the population. Undoubtedly they are an important factor in the promotion of a homogeneous national taste and spirit.

A certain large group of programs may be classified as educational. This group includes lectures on pure and applied sciences, the arts, industries, (and, in general, such material as may be included in university lecture courses open to the public). It also includes political discussions, and has been held by some to embrace news reports as well.

Such programs undoubtedly are of great value in welding a nation together into a well-defined political unit and also, under favorable conditions, in placing its culture and viewpoint directly before other nations. The spread of knowledge is furthered by such programs which, it may be mentioned, are as well a powerful agency toward the creation of a single language used and pronounced identically in all portions of the country which is served. It is not necessary to point out further the essentially unifying characteristic of such programs for any nation.

A third group of programs which have both national and international aspects are those which are religious in character. Such programs include both sectarian and non-sectarian presentations. In most instances, the latter have been found to be more suitable for national distribution. Tolerance and mutual understanding are promoted by such programs, and it is therefore believed that presentations of religion, ethics, and morals are an important and permanent portion of any well-balanced plan for national broadcasting.

There are in addition numerous non-technical problems involved in connection with the national aspects of broadcasting. Many complicated questions of national policy arise. To what extent shall governments scrutinize or even censor programs in times of peace? In what measure shall the doctrine of the "freedom of the press" be held to apply to national broadcasting? On what basis, if any, shall this great system of mass communication be restricted in scope or choice of material by other forces than the dictates of common sense and good taste?

These and other questions intimately connected with the influence of broadcasting upon national solidarity affect all national legislation and broadcasting procedure and are complicated problems to which only the future evolution of mankind and of broadcasting can give the answers.

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RADIO BROADCASTING TRANSMITTERS AND RELATED TRANSMISSION PHENOMENA*

By

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THIS paper is a brief discussion of recent developments in American practice concerning radio broadcasting transmitters. Descriptive material and photographs pertaining to several new commercial transmitting equipments are included. Reference is also made to the more important aspects of the related transmission problem. On account of the scope of the subject, the treatment is necessarily superficial, but it may serve to indicate the present status of the transmitter art and its relative position with respect to the industry as a whole. A short bibliography containing some of the more important recent contributions to the subject is attached as an appendix, to which reference may be had for more detailed information.

RADIO TRANSMITTERS

The radio transmitter is essentially a focal point in the present-day broadcasting system, since upon it the program circuits converge and from it the radio distribution network emanates. For this reason, the requirements which have been imposed on transmitting apparatus are extremely rigorous, and all phases of transmitter performance have been subjected to the most careful scrutiny. Under these stimulating influences, the last few years have brought about some very noteworthy advances in this portion of the broadcasting field.

As long as music and entertainment continue to hold a prominent place on broadcasting programs, fidelity of transmission will probably remain the most sought-for characteristic, not only for the radio transmitter itself, but for all of the apparatus units in the system. A very high standard of performance has now been attained in this respect. Fig. 1, below, shows the overall frequency-response characteristic of a new type 50-kw equipment, the first of which has gone into service at one of the leading American broadcasting stations within the past few months. It will be noted that this characteristic is substantially flat between 30 and 10,000 cycles. The greatest departure from the horizontal line which is the ideal characteristic is less than 1 db. The frequency discrimination which this represents is of such a low order that it probably could not be detected in ordinary listening tests, even by a skilled musician.

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Another recognized prerequisite to a high degree of fidelity is exact proportionality between audio input and sideband output. Increased emphasis on accurate reproduction has recently led to the introduction of improved technique for checking this important characteristic under dynamic conditions. The method employed consists of impressing a pure sine-wave input on the transmitter at various frequencies throughout the audio range and subjecting the output of a straight-line rectifier to harmonic analysis. One type of harmonic analyzer which has been used with excellent results is that due to Wegel and Moore.[1] This device produces a photographic record, an example of which is shown in Fig. 2. Measurements of this type are of particular importance under present conditions since current American practice is tending toward the extensive use of transmitters in which modulation is accomplished at relatively low power levels and the required power output is obtained by means of subsequent stages amplifying modulated radio-frequency power. Such amplifying stages are susceptible of serious amplitude distortion unless the conditions under which the tubes operate (direct plate and grid voltages and impedance of the connected load) are carefully predetermined. For this

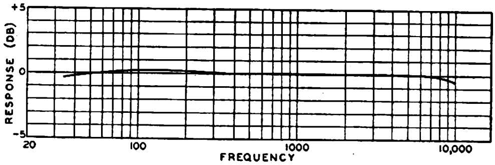

Fig. 1—Frequency-response characteristic of Western Electric 7-A (50-kw) radio transmitter.

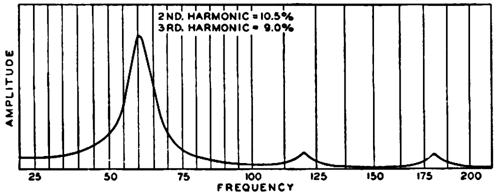

Fig. 2—Harmonic analyzer graph indicating overloading (2nd harmonic, 10.5 per cent; 3rd harmonic, 9 per cent).

[1] R. L. Wegel and C. R. Moore, "An Electrical Frequency Analyzer," *Bell Syst. Tech. Jour.*, p. 299–323, April, 1924.

purpose, the harmonic analyzer has proved to be invaluable. Through its use, commercial transmitters are now available in which, at the working upper limit of modulation, the harmonics generated are not greater than 5 per cent.

The attainment of such high standards for fidelity leaves little opportunity for progress, and it is improbable that significant advances in this direction will be made in the near future. Accordingly, in continuing their efforts toward further improvements in broadcasting service, transmitter engineers have been led to divert their attention to the problem of rendering less conspicuous and objectionable the background of noise and interference which, in the past, has so seriously impaired the artistic effect of programs except in the immediate vicinity of transmitting stations. This is the principal justification for the present movement toward higher power outputs for broadcasting stations. It has also resulted in increased emphasis on the maintenance of a high average degree of modulation, a development which is rapidly bringing about a very perceptible improvement in general broadcasting conditions.

The degree of modulation of the carrier in a radio telephone transmitter is a somewhat intangible factor which necessarily varies rapidly through wide limits during the rendition of a program. With every transmitter, however, there is a definite modulation limit which is a characteristic of the design and which cannot be exceeded without bringing about serious distortion. This limit is an important performance index which, for lack of a better name, has been called "modulation capability." The modulation capability of a transmitter may be defined as the maximum degee of modulation (expressed as a percentage) that is possible without appreciable distortion, employing a single-frequency sine-wave input and using a straight-line rectifier coupled to the antenna in conjunction with an oscillograph or harmonic analyzer to indicate the character of the output.

For a number of reasons, some technical and some economic, many of the broadcasting transmitters in use have been so constructed that overloading of the audio power stage with consequent distortion occurs whenever the degree of modulation exceeds approximately 50 per cent. The usual practice in placing broadcasting transmitters into service consists of determining, by means of a suitable vacuum-tube voltmeter or other "volume indicator," the audio level at the input of the set for which distortion becomes evident. The average operating level is then established at a suitably lower value, frequently 6–10 db. Recently, by modulating at low power levels, transmitters have been produced which are capable of 100 per cent modulation without noteworthy distortion. It is obvious that, if a transmitter of this latter type is employed and the same margin is observed in determining the average audio input level, the resulting sidebands will be twice the amplitude of those produced by a transmitter whose modulation capability is only 50 per cent. To produce equivalent sidebands with a transmitter capable of but 50 per cent modulation requires that the carrier amplitude be doubled or the power output multiplied by four. In other words, insofar as signal-to-noise ratio is concerned, which is the factor that usually determines the coverage of a broadcasting station, the increase in modulation capability mentioned results in an improvement that in the older type of apparatus could only be had by quadrupling the rated output of the transmitter. From the coverage standpoint, the night range of a given station can be approximately doubled in this manner. Since this is accomplished without increase in the carrier power, the outlying zone in which the station may produce serious beatnote interference with others assigned to the same channel will not be extended. Accordingly, the use of transmitters capable of a high degree of modulation is a notable contribution toward the more effective utilization of the medium, which is the outstanding technical problem in American broadcasting today.

Another important factor, from the standpoint of intensive development of the available frequency band, is frequency maintenance. In a system involving so many stations as are now operating in the United States, accurate maintenance of the assigned frequencies presents a very difficult problem. The maximum deviation permitted by the existing government regulations (± 500 cycles) is somewhat beyond the capabilities of the ordinary wavemeter and difficulty has been experienced in obtaining a satisfactory substitute. In the absence of adequate frequency control apparatus, very serious beatnote interference has been of frequent occurrence. During the past year, however, considerable improvement has been brought about by the extensive adoption of piezo-electric reference oscillators and automatic piezo-electric control. Equipment for the latter purpose capable of a relatively high standard of performance is now being offered commercially and it is probable that apparatus of this type will be installed in the near future by the majority of stations. Its use is expected to avoid entirely heterodyne interference on the "cleared" channels, where the beatnotes are those produced between the carriers of stations having adjoining frequency assignments. There is also reason to believe that the general adoption of such apparatus will materially improve conditions on the "shared" channels, each of which is occupied by several stations located at suitable distances, provided the assigned frequencies can be maintained with sufficient accuracy to preclude the reproduction of audible beats or other objectionable interference effects.

This problem of "synchronization," or preferably "common frequency operation," is beginning to receive considerable attention from all factors in the broadcasting industry. It promises important contributions in at least two directions:

(1) Improvements in the coverage of a common service area by two or more stations all broadcasting the *same* program;

(2) The attainment of minimum geographical spacings between stations operating on the same nominal frequency and broadcasting *different* programs.

The degree of frequency maintenance required for these two cases is apparently quite different. For case (1), the evidence indicates that very rigorous requirements must prevail. The most successful operations of this type have employed wire lines connecting the stations for the transmission of a base frequency from which the carriers were derived by means of harmonic generators. For case (2), however, there is reason to believe that comparatively wide limits will suffice.

Experience has shown that if the entertainment value of a program is not to be seriously impaired by interference, the ratio of wanted to unwanted carrier at the receiving point, in terms of field intensity, must be at least 100:1. From a relative interference standpoint, the significant factors are the wanted sidebands, the unwanted sidebands and the unwanted carrier, each of which produces a component in the detector output by interaction with the wanted carrier. With equal modulation at both stations, which is one of the conditions assumed, the ratio of the audio components due to the sidebands will, in general, be approximately the same as that between the carriers, or 100:1, representing a difference in level of 40 db. Due to the frequency difference between carriers, demodulation of one of the unwanted sidebands will result in the original signal with each of its elements shifted upward in pitch by an amount corresponding to this difference, while the other sideband will produce a signal which is similarly displaced in the reverse direction. The interfering signal may be badly garbled, therefore, but its disturbing effects insofar as enjoyment of the program is concerned will be substantially unaffected. The beatnote, which results from the interaction of the unwanted and wanted carriers, will be 6–10 db above this sideband interference level if average practice, as previously described, is followed. From this analysis, it appears that if the beatnote can be held to a value below the lowest frequency which it is desired to transmit and if one of the circuit elements of the reproducing system can be designed to provide some 10 db discrimination against the beat frequency, interference due to the latter can be so subordinated that the service areas of the stations involved will be defined by the limiting condition assumed for sideband interference, or a 100:1 ratio between carrier field intensities. Under these circumstances, no beatnote interference will be experienced in those areas where reasonably good service can be given. In adjoining regions, where the carrier ratio is less than 100:1, beatnote interference may continue to be observed but is of no importance since satisfactory reception in such areas is precluded by the sideband interference.

To meet the requirements outlined, it is probable that ultimately frequencies will have to be maintained to approximately 10 cycles, which would result in a maximum beatnote near the lower limit of aural frequency response. Such precision seems hardly necessary, however, under the conditions existing at the present moment. Almost all loud speakers now commercially available discriminate notably against frequencies below 100 cycles. A material improvement in beatnote conditions could probably be brought about, therefore, by the adoption of automatic control apparatus capable of maintaining the assigned frequencies to ±50 cycles. Such performance is within the capabilities of the piezo-electric apparatus now available. Under the circumstances it is expected that considerable progress will be made in this direction during the coming year.

The foregoing considerations lead to the formulation of an important system requirement affecting receiving apparatus, which in this case includes both the radio receiver proper and the loud speaker. In a system involving a relatively large number of stations assigned to cleared and shared channels at 10-kc intervals, such as exists in the United States, beatnote interference in the form of components at approximately zero cycles and at 10 kc is an inherent characteristic. If a maximum frequency deviation of ±10 cycles is accepted as the ultimate limit, in order to avoid such interference the receiving apparatus must be so designed that at frequencies below 20 cycles and above 9980 cycles there will be introduced sufficient attenuation to suppress effectively the beatnotes likely to be encountered under any practical operating condition. Developed in this manner the proposition is more or less self-evident, but due to the rapidity with which the audio spectrum of broadcasting apparatus is being extended, some emphasis on the matter seems desirable.

Still another factor of importance from a system standpoint is control of radio harmonics. Spurious radiation of all types is inimical to intensive development and must be avoided. The harmonic problem presents unusual difficulties since efficiency requires that the tubes in the final power-amplifier stage be used in such a manner that relatively large harmonic voltages are impressed on the output circuit, yet the harmonic power radiated must be held to an extremely small amount. A measure of the purity of wave form required may be gained from the fact that a 5-kw transmitter operating on a good antenna is capable of establishing an electromagnetic field of approximately 0.5 v per meter at a distance of one mile. Under the circumstances, a harmonic of 0.1 per cent represents a field intensity of 500 μv per meter at the same distance. Acceptable service in many areas is being obtained with field intensities of this order of magnitude. To bring the interfering field down to the static level would probably require reduction of harmonics to 0.01 per cent or less. From an apparatus standpoint, such performance represents a very difficult problem and it is questionable if it can be justified at the present time. Practice on this point is still in a state of flux, but there is reason to believe that some intermediate value, such as 0.05 per cent, will prove to be the proper solution, and will be applied to all broadcasting stations in the near future.

One circumstance that has undoubtedly contributed to the delay in formulating definite requirements concerning the control of harmonics has been the difficulty of obtaining suitable apparatus for the evaluation of such components in quantitative terms. Field strength measuring sets have recently been made commercially available, however, which are capable of covering the necessary range in frequency and intensity. A photograph of one of these sets is shown in Fig. 3. It consists essentially of a sensitive, stable superheterodyne receiver incorporating a calibrated attenuator at the input of the intermediate-frequency amplifier and a supplementary radio-frequency oscillator from which a voltage of the frequency of the station under measurement can be introduced in the antenna circuit. The operating characteristics of such an instrument have been described by Friis and Bruce.[2] By means of a series of removable loops and coils, the set shown is capable of measuring field strengths ranging from approximately 0.01 to 7000 mv per meter throughout the band 250 to 6000 kc. Apparatus of this type is now in use by the radio inspection division of the Department of Commerce.

In the light of this discussion of present trends in transmitter development, a brief description of some recent transmitting equipments may be of interest. A particularly noteworthy example of current practice is the 50-kw Western Electric transmitter, one of which has been placed in service within the past few months by the Crosley Radio Corporation at Mason, Ohio. Views of this equipment are shown in Figs. 4, 5, 6, 7, and 8. The transmitter proper is shown in Fig. 4. As will be seen, it consists of seven panel units with a screen enclosure in the rear. The first unit on the left is the

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[2] H. T. Friis and E. Bruce, "A Radio Field-Strength Measuring System for Frequencies up to Forty Megacycles," Proc. I. R. E., 14, 507—519; August, 1926.

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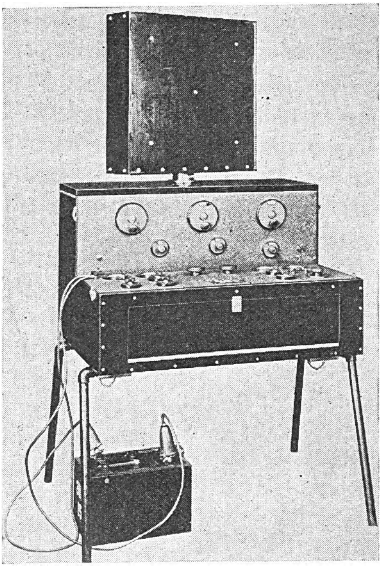

Fig. 3—Commercial field-strength measuring set. Range: 250—6000 kc, 0.01—7000 mv per meter.

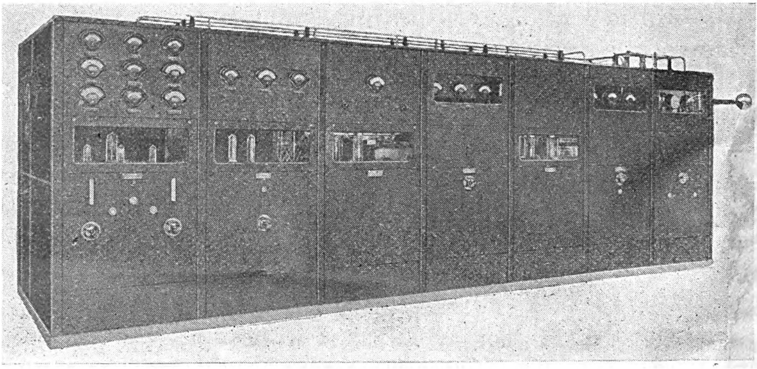

Fig. 4—Western Electric 7-A (50-kw) radio transmitter. Oscillator-amplifier assembly.

oscillator-modulator. This is essentially a low-power transmitter capable of an output of 50 watts and 100 per cent modulation. It is followed by three push-pull stages amplifying modulated radio-frequency power. The first power-amplifier stage, which employs two 250-watt tubes, occupies the second unit. The tubes for the second power stage, which are water cooled, and the associated tuned output circuit are contained in the third and fourth units, respectively. The final power stage, incorporating six water-cooled tubes each capable of a peak output of approximately 40 kw, occupies the fifth unit. The last two panels constitute the front of an electrically screened enclosure housing the output circuits for this latter stage. All of the panels are aluminum covered with several coats of black lacquer grained by rubbing with abrasive paper. A full complement of meters is provided, the cases of which are either grounded or mounted behind glass for the protection of the operating personnel. In designing the equipment, special consideration has been given to safety. Access to the apparatus in the rear of the panels can be had only through the door on the left which is held closed by a bolt operated by the handwheel shown. The rotation of this wheel opens the transmitter control circuits putting the equipment out of operation. It then grounds the high-voltage supply busses and finally withdraws the bolt. As an additional precaution a manually operated disconnect switch for the main power supply is provided just inside the gate which can be opened on entering. Access to some of the tubes is had by opening the glass windows in the various panels, but these are secured by electrically operated latches unless the wheel is in the grounded position. Door switches are provided in the control circuits which prevent the transmitter from being placed in operation unless all doors and windows are closed.

The power panel and rectifier assembly is shown in Fig. 5. The general arrangement corresponds to that of the transmitter proper and similar safety features are provided. In the power panel, which is on the left, are centralized the necessary power distribution and control facilities. The equipment requires a 3-phase input of approximately 250 kw at 440 volts. The control arrangement is such that the transmitter can be started and stopped by means of a single set of push buttons, the various circuits being energized in proper sequence by means of suitable relays and contactors. The central unit is a three-phase half-wave rectifier supplying power at 1600 volts to the plates of the air-cooled tubes. The six-tube rectifier on the right supplies plate power at 17000 volts for the water-cooled tubes. The filament and plate transformers and smoothing filter for the latter are located in the power room on the floor below. The filter consists of two units, one for each side of the push-pull circuit, employing a 6-μf condenser and a 12-henry inductance. Two 24-volt, 550-ampere direct generators (one a spare) supply power to the filament circuits. These machines are slot wound and employ composition brushes, a filter consisting of a 1-mh inductance and four 1000-μf electrolytic condensers being used to suppress commutator and slot ripples. Grid bias voltages are obtained from a 2-kw, 300-volt unit, which is also installed in duplicate. The only other rotating apparatus is that associated with the water-cooling system. The tubes are cooled by means of distilled water which is conducted to the anodes of the amplifier tubes through insulating hose coils. The total heat transferred by the cooling water is approximately 175 kw. A flow of 75 gallons per minute is maintained. Four 56-in. by 58-in. radiator units are employed, each consisting of a bank of copper tubes with spiral fins over which air is blown by a 37-in. fan. Ample radiator capacity is provided to maintain the water below 180 deg. F under all atmospheric conditions.

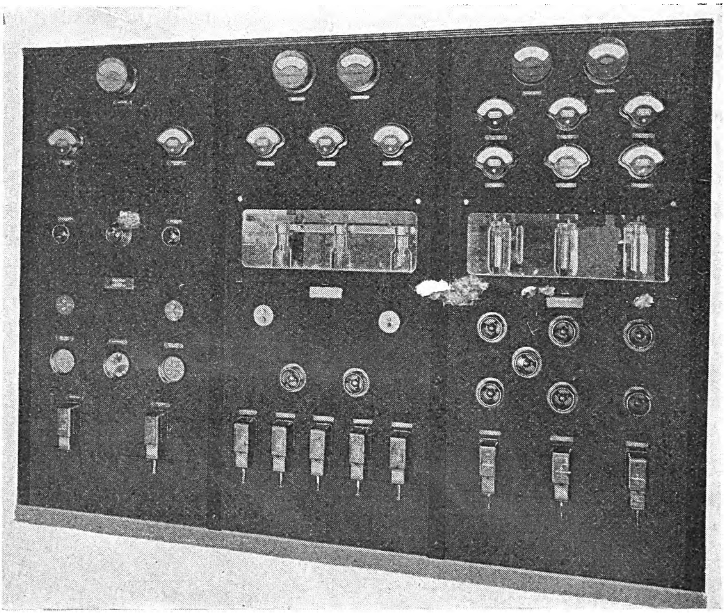

Fig. 5—Power panel and rectifier assembly for 50-kw radio transmitter.

To promote antenna efficiency and to reduce the intensity of the electric field in the station building, the equipment is arranged to deliver its output to the antenna through a radio-frequency transmission line approximately 500 ft. long. The line is balanced to ground and is designed for a characteristic impedance of 600 ohms. The antenna coupling and tuning unit is shown in Fig. 6. It is intended for installation in a small building with a grounded copper roof located at the base of the antenna downlead. It consists of two tuned circuits, each housed in separate shielded compartments. In the photograph the doors and two of the screen panels have been removed to show the interior arrangement. The line is terminated by the parallel tuned circuit on the left which is inductively coupled to the antenna circuit to preserve an approximate balance to ground. The antenna is tuned by means of the series condenser and coil shown on the right. Accurate adjustment of the inductance of the coil is provided for by means of a short-circuited single-turn secondary which is located within the coil and arranged so that it can be rotated through approximately 90 deg. by the motor mounted on the floor beneath. The latter may be controlled from the operating room by a reversing switch placed on the right-hand panel of the transmitter assembly.

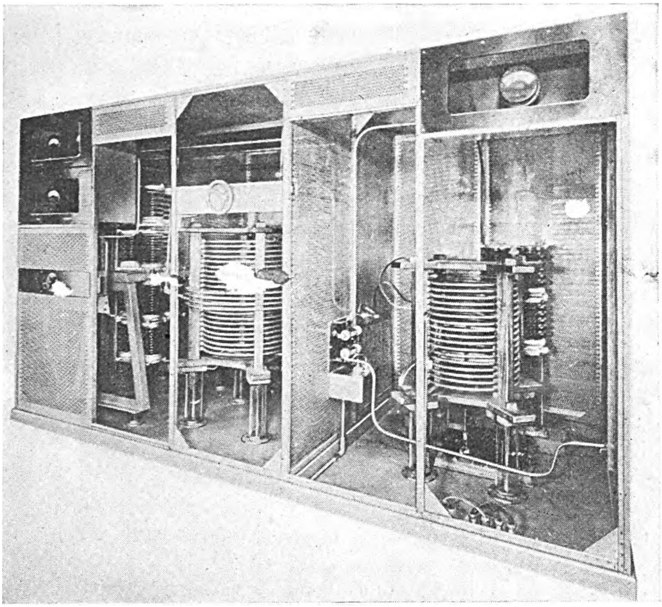

Fig. 6—Antenna coupling and tuning unit for 50-kw radio transmitter.

A polyphase position indicator is provided to indicate the angle and movement of the secondary. The direct-current circuit of the thermal ammeter in the antenna circuit is also carried back to a bracket-mounted instrument on the end of the transmitter. These facilities permit the antenna tuning to be checked and adjustments made to compensate for minor variations in antenna conditions without leaving the operating room.

Another feature of interest is the artificial antenna shown in Fig. 7. This unit is essentially a 600-ohm non-inductive resistance capable of dissipating approximately 75 kw which can be connected to the output circuit of the final power amplifier stage in place of the transmission line. The heat dissipating elements consist of a series of woven wire grids mounted in the units at the top of the framework. The resistance of these grids is substantially independent of frequency, but the combination presents a slight inductive reactance which is compensated for by means of the condenser and coil combination shown. These elements are inserted into the circuit symmetrically in order to maintain an approximate balance to ground. The structure is completely shielded and is fitted with safety door and grounding switches similar to those already described.

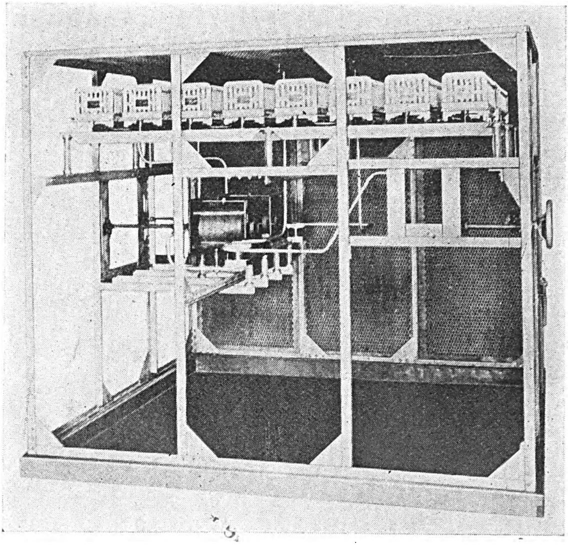

Fig. 7—Artificial antenna for 50-kw radio transmitter.

The piezo-electric crystal mounting and temperature-control apparatus which is a part of the oscillator-modulator unit is shown in Fig. 8, dismantled to facilitate inspection. The quartz plates employed are approximately one and a quarter inches square and are cut parallel to one of the faces of the natural rock crystal. This plate is mounted between two lapped metal plates and covered with a porcelain cap carrying a terminal to which the upper electrode is connected by means of a short section of metal foil. The mounted crystal is supported by a brass block, through the center of which extends a spiral bimetallic thermostat. The top of the block is also lapped and the crystal mounting is secured to it by means of the four springs shown. The heating element consists of a winding of resistance wire inserted in the block concentric with the thermostat. The assembly is mounted in a thermally insulated box, shown on its side in the photograph.¶ Two of these units are provided, one located on each side of the oscillator-modulator unit directly below the window. A detachable handle for adjusting the contacts of the thermostat and a suitable thermometer extend through the box to the front of the panel. The brass mounting block is provided with a groove to receive the bulb of the thermometer. The thermostat does not operate directly in the heater circuit but controls the grid bias of a vacuum tube in the plate circuit of which a suitable relay is placed. The quartz plates are ground to oscillate at the assigned frequency at approximately 50 deg. C, and the final adjustment is made by varying the operating temperature. The temperature coefficient of the plates varies from 30 to 100 parts in a million per deg. C. The degree of constancy attained necessarily depends on the diligence of the operating personnel. With proper maintenance the maximum deviation has been held to ± 30 cycles for long periods of time.

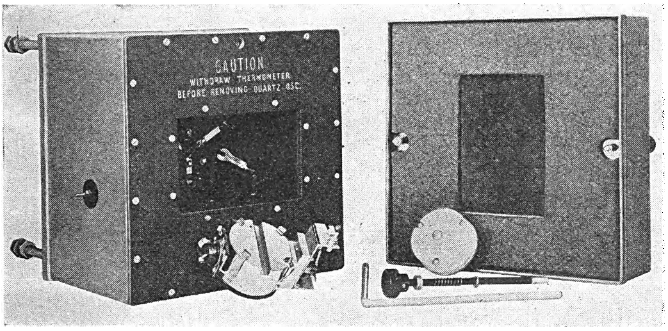

Fig. 8—Piezo-electric crystal mounting and temperature control apparatus.

A simplified circuit schematic is shown in Fig. 9. Features of the electrical design are the modulation system, the push-pull amplifier

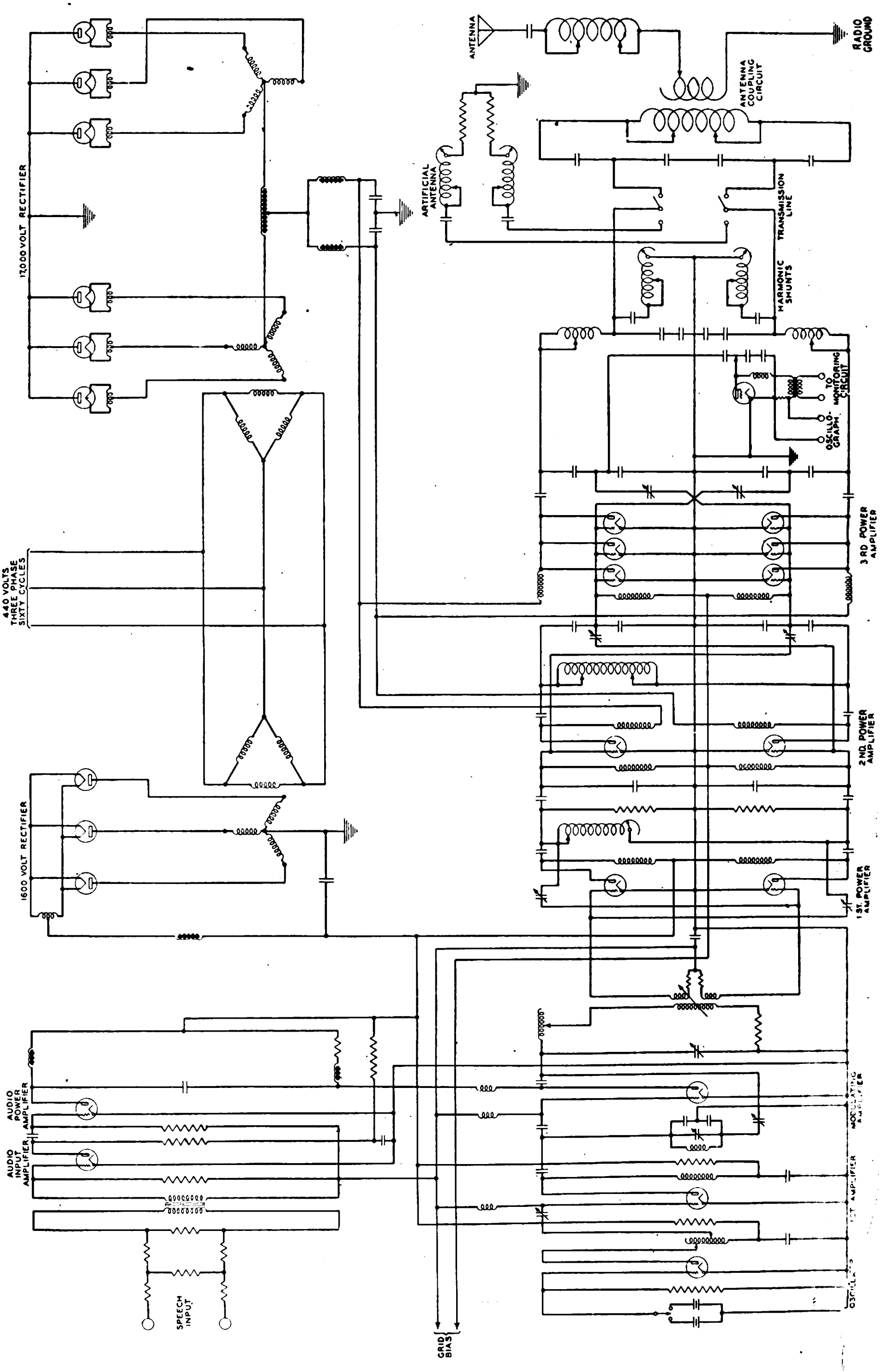
ANTENNA
RADIO GROUND
ANTENNA COUPLING CIRCUIT
ARTIFICIAL ANTENNA
TRANSMISSION LINE
HARMONIC SHUNTS
TO MONITORING CIRCUIT
OSCILLOGRAPH
17,000 VOLT RECTIFIER
440 VOLTS THREE PHASE SIXTY CYCLES
3 RD POWER AMPLIFIER
2 ND. POWER AMPLIFIER
1 ST. POWER AMPLIFIER
1600 VOLT RECTIFIER
AUDIO POWER AMPLIFIER
AUDIO INPUT AMPLIFIER
SPEECH INPUT
GRID BIAS
MODULATING AMPLIFIER
1 ST. AMPLIFIER
OSCILLATOR

Fig. 9—Simplified circuit schematic of

stages with cross neutralization, the capacity coupling arrangement used to facilitate control of parasitic oscillations, and the provisions for the suppression of harmonics. The modulating amplifier is a 50-watt tube operating at 750 volts. The audio power stage employs a 250-watt tube at 1500 volts. In this manner, ample audio-frequency voltage and power are provided to effect complete modulation without distortion in the audio tube. With so powerful an equipment, the suppression of radio-frequency harmonics to a satisfactory degree becomes a difficult problem. The push-pull circuits, capacity coupling, three tuned circuits in cascade, shielding of all coils, and the two tuned shunts adjusted to the second harmonic which are connected between each side of the transmission line and ground all contribute to superior performance in this respect. The amplitude of the harmonics radiated, as determined by field strength measurements, is less than 0.03 per cent.

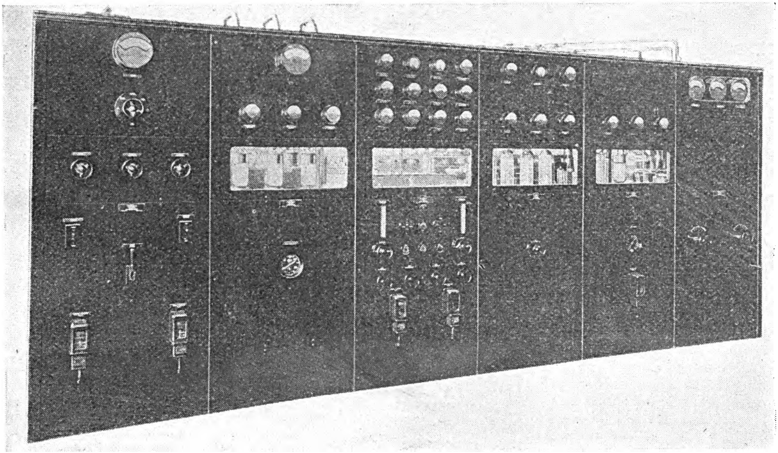

Fig. 10—Panel assembly for Western Electric 5-C (5-kw) radio transmitter.

A 5-kw equipment of similar general design is shown in Figs. 10 and 11. It consists of six units: A power panel, a 10,000-volt rectifier for the water-cooled tubes, a piezo-electric oscillator unit, an intermediate amplifier unit, a power amplifier unit employing two 10-kw tubes, and an output unit. An air-cooled transformer for the rectifier, the associated filter, and an artificial antenna are assembled in a screened enclosure in the rear of the panels. Three motor-generator sets are provided to supply filament power, grid bias, and plate power for the air-cooled tubes. A 3-phase power input of 30 kw at

220 volts is required. The equipment is capable of fidelity in transmission comparable with that of the 50-kw unit. The amplitude of the harmonics radiated is held to approximately 0.2 per cent.

A 1-kw equipment of the same type is shown in Figs. 12 and 13. It involves only two panels, a piezo-electric oscillator unit and an amplifier unit. The final power stage employs a 4-kw water-cooled tube. Two motor generators are used, one supplying 24 volts and 250 volts for filaments and grid bias, the other 2,000 volts and 4,000 volts for the plates of the air-cooled and water-cooled tubes, respectively. A power input of 10 kw is required.

Radio Transmission Phenomena

Radio transmission phenomena in the broadcasting band have been given considerable study, and the general nature of the effects likely to be encountered are fairly well understood. Important contributions have been made by Bown and Gillett, by Bown, Martin, and Potter, by Goldsmith, and by Espenschied.[3] The second paper referred to is particularly noteworthy on account of the insight which it affords into the complexities of the process of transmission and the evidence which it presents concerning the injurious effects of frequency modulation. The latter has not yet fully received the attention which it deserves; many otherwise well designed transmitters are still in operation that are subject to frequency changes of the order of ±1000 cycles during modulation. This condition is not only conducive to impaired fidelity at moderately distant receiving points,

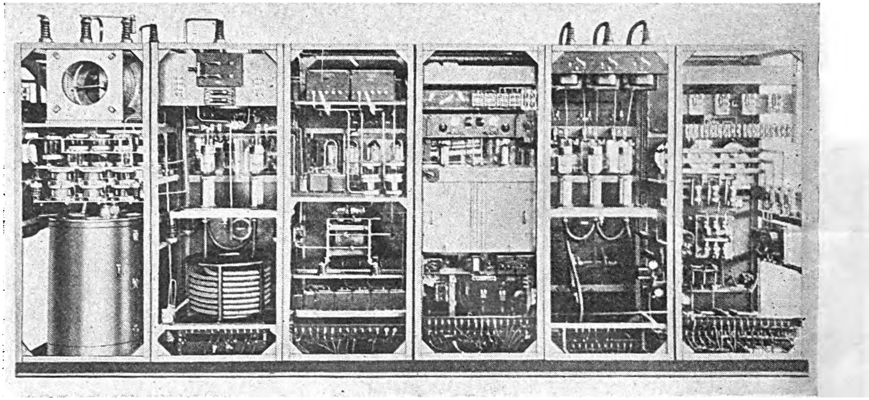

Fig. 11—Rear view of panels in 5-C radio transmitter.

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[3] See attached list of references.

but it increases interference and precludes successful common frequency operation. Fortunately, the use of automatic frequency control apparatus in its present form is effective in minimizing this effect as well as in limiting frequency variations of much longer period. It is probable, therefore, that with the more general use of automatic piezoelectric control, this matter will rapidly cease to be a problem.

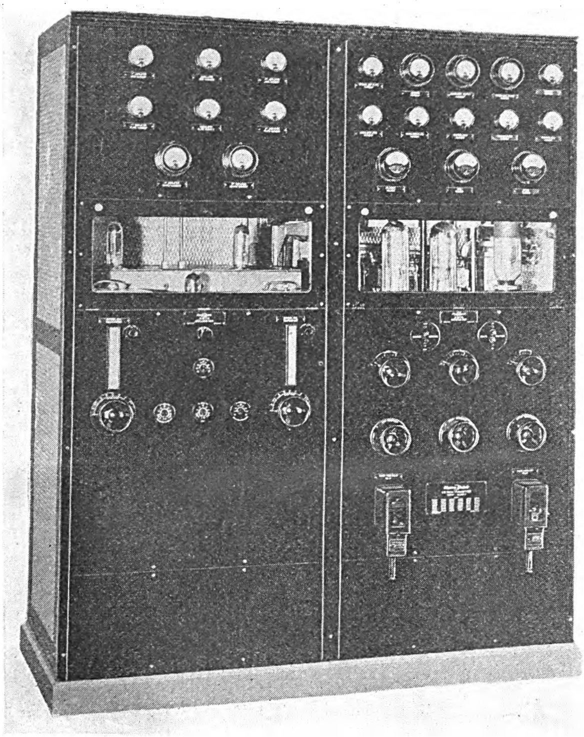

Fig. 12—Western Electric 6-B (1-kw) radio transmitter.

As might be expected, the attention being given to intensive development has materially stimulated interest in transmission. There is a very evident need for much information of a more quantitative nature than is now available. Data concerning attenuation over city and rural areas as a function of frequency, suitable separations between stations of various powers operating on a common carrier frequency, allowable distances between transmitting stations and nearby populous communities, relative day and night ranges, relative summer and winter ranges, time of the day and season of the year at which the transition occurs, and other questions of a similar nature have become of great practical importance. The problem is rendered particularly difficult by the range in climatic, topographic, and cultural conditions which exist in the United States. Under the circumstances, there are excellent opportunities for important work in this field.

A significant tendency disclosed by recent measurement work in a number of city areas is public acceptance of and demand for field intensities which a few years ago would have been considered objectionably high. For some time it has been more or less generally agreed that a field intensity of 10 mv per meter would afford a satisfactory high-grade broadcasting service. Recently, however, in spite of increased effectiveness due to higher degrees of modulation and in spite of continued improvement in the sensitivity of commercial receiving sets, stations establishing field strengths of 10–15 mv per meter have been greatly handicapped in competing with others capable of produc-

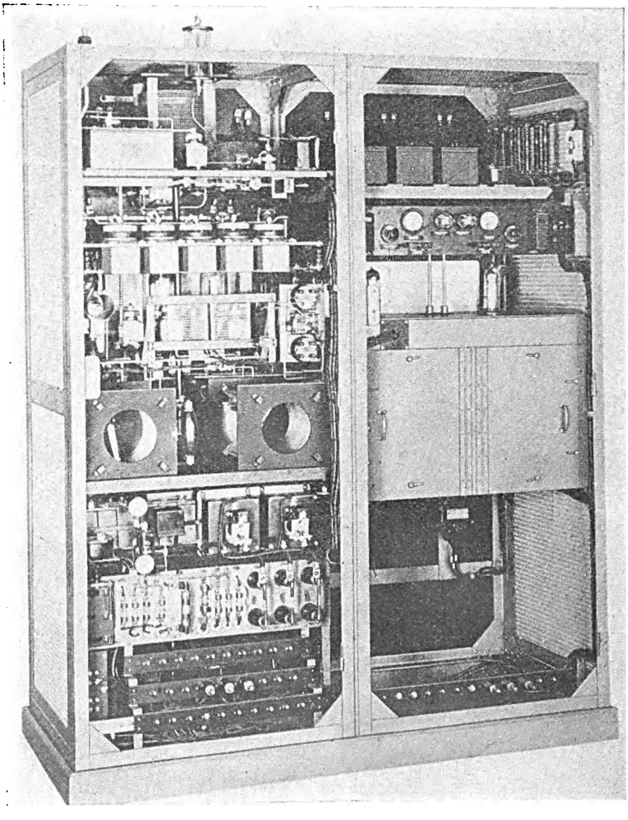

Fig. 13—Rear view of 6-B radio transmitter.

ing 30–50 mv per meter in the same areas. In several densely populated districts measurements have disclosed field intensities of 300–500 mv per meter without any noteworthy number of complaints provided the programs were of a high character. There is little to indicate whether this tendency is the result of a decreased interest in distant stations, a desire for higher standards in reproduction involving lower noise levels, or a combination of these factors with others, but it is evidently a matter which must be given careful consideration in engineering future installations.

It is interesting to contrast this situation with that existing in some of the large rural districts as exemplified by the recent survey of conditions in the Middle West by Jansky.[4] Here over large areas acceptable service is being obtained with field strengths of 50 and 100 μv per meter. Giving due consideration to the difference in noise levels, which is undoubtedly a factor of great significance, such a discrepancy can only be reconciled on the basis of a vast difference in service standards. That such conditions will be allowed to continue for any considerable period of time is very doubtful. This is further evidence indicating that the movement toward more powerful stations is technically sound.

One phase of the transmission problem which deserves increased attention is antenna performance and design. It is an interesting circumstance that while the accurate rating of broadcasting stations s a matter of great practical concern to the industry, to date conideration has been confined to the power delivered to the antenna. Variations in the efficiency of the latter have been almost entirely negected in spite of the fact that, due to this cause, the power actually radiated can be shown to vary through a range of four to one, or greater. There is little doubt that stations should be rated, either directly or indirectly, in terms of field intensity measurements. That such a system of rating has not already been put into effect is probably due to the lack of suitable measuring apparatus. With such quipment now available, rapid progress in this direction is expected.

An interesting feature of current American practice with respect o broadcasting antennas is a definite tendency toward the use of igher supporting structures. For the past few years, most of the towers erected have been from 150 to 225 ft. in height. Several of the more recent stations are employing 300-ft. towers, and it is not improbable that some 400-ft. structures will be put up in the near future. Since the natural frequency of grounded steel towers of these dimensions falls in the broadcasting band and may approximate the assigned operating frequency, low-capacity porcelain insulators are inserted at the base. The latter effect a considerable increase in the natural frequency of the towers and preclude serious distortion in the field intensity pattern due to heavy induced currents in the steel. The antennas themselves are of such dimensions that the current antinode is positioned well up on the vertical section. The effect is to concentrate the radiated power along the ground plane and to increase materially the field intensity in the local service area. Such antenna systems promise a better economic balance between the investment for generating modulated radio-frequency power and that for radiating it.

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[4] See attached list of references.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 11 *November, 1929*

THE NATIONAL BROADCASTING COMPANY, A TECHNICAL ORGANIZATION FOR BROADCASTING*

BY

JULIUS WEINBERGER

(Engineer in Charge, Research Department, Radio Corporation of America, New York, N. Y.)

RADIO broadcasting on a nationwide scale commenced in the United States approximately in 1920. Late that year, and during the years immediately following, the Westinghouse Electric and Manufacturing Company and General Electric Company set up broadcasting stations at several of their manufacturing plants and began to give programs of speeches and music, at first of an experimental nature, but soon of a high quality and systematic character.

In July, 1922, the American Telephone and Telegraph Company, anxious to study the possibilities of this new medium, just as it had previously studied the possibilities of radiotelephony in private or point-to-point communication, inaugurated Station WEAF in New York City and began to give a regular series of programs.

The free and unrestricted growth of broadcasting which soon followed was characteristic of the United States. Begun at first for the purpose of encouraging the sale of radio equipment to the general public by radio manufacturing companies, the potentialities of this means of organized mass communication were soon realized by newspapers, commercial houses, banks, theatrical organizations, schools, churches, and private individuals. Anyone who felt that he had a message that would be of interest to the public set up a broadcasting station, and by August, 1924, a total of 1105 stations were in operation. The cycle of development soon became established, with the broadcaster first sending out programs, then the community expressing its interest and support by equipping its homes with radio receivers, back again to the broadcaster increasing his efforts by way of better programs and service in keeping with his increasing audience, followed by still more and better receiving facilities.

During the early part of this period of growth, little cognizance was taken of the cost of broadcasting. Those operating stations felt, for a time, that the good will or educational value of their work justified the expense. As time went on and program standards gradually rose, the high cost of broadcasting began to have its economic effect. By 1926 the number of stations had dropped to 533, or less than half the number in operation in 1924. "Who shall pay for broadcasting?" became the question of outstanding importance. Other nations, following us in the matter of inaugurating a broadcasting service, had applied strict licensing laws regarding the installation of radio receivers and were exacting licensing fees as a means of defraying the cost of broadcasting. The listeners of many countries were made to pay the bill, and they do so today. However, it would have been impossible to establish such a system in America, due partly to the inherent nature of the people and partly to the widely diversified control under which broadcasting had grown up.

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Aside from the economic burden of broadcasting, there was a second factor which led to the early mortality of stations. This was the paucity of really suitable broadcasting program material. Stations in smaller centers, which had begun with an apparently ample supply of available talent in their community, soon found themselves nearing the exhaustion of their local program resources. They were hard pressed for good programs whereby to maintain the interest of their audiences.

On the other hand, the programs of such stations as WJZ and WEAF, located in or near New York City,—the amusement center of the nation and the scene of its most important events—were steadily improved and extended in scope. It was soon apparent that such stations could develop and maintain a program standard of greater breadth, interest, and value than the stations less fortunately situated.

After the original wonder of broadcasting wore away, the companies engaged in the manufacture and sale of radio receiving equipment soon recognized that the demand of listeners was for programs of greater and greater interest and importance. It became obvious that if radio broadcasting was to perform a great public service in this country, it would be necessary to develop chains of stations, linked together by telephone wires, in a system which would be capable of picking up the best program material anywhere in America and broadcasting it everywhere. This was not only necessary in order to render a public service, but it was necessary if the market for radio receiving sets was to be extended and enlarged.

FORMATION OF THE NATIONAL BROADCASTING COMPANY

This company was organized to meet the two obvious necessities of advanced broadcasting, namely, (1) adequate financial strength, *ultimately self-sustaining;* and (2) adequate program material, with an organization and equipment capable of coordinating and presenting this material in the most desirable manner.

A most important step in the direction of making broadcasting self-sustaining had been taken by the management of station WEAF of New York City, then the American Telephone and Telegraph Company. This station originated the practice of selling its services to commercial organizations for advertising purposes. It was the intent that an advertiser would provide a program of interest to the radio listeners, incidentally mentioning his product, thus creating useful goodwill toward his wares, and pay the station for its time and facilities. At first, many of those interested in the radio field felt that the introduction of this type of broadcasting would ruin the industry. However, it has proved to be the only foundation from which a self-sustaining broadcasting service in the United States can be built up, as practiced today under the term, "sponsored programs."

In addition, this station had formed a telephone wire network linking together some nineteen broadcasting stations, operating under a wide variety of managements in seventeen different cities, known as the "red network." The programs of WEAF were supplied, for part of the day, to these so-called "outlet stations." The total network comprised at that time approximately 3600 miles of special telephone lines.

One of the first steps of the newly formed National Broadcasting Company was to purchase this station and to lease its existing telephone line network, the American Telephone and Telegraph Company then retiring from the broadcasting field.

In the meantime, the organizations which had formed the National Broadcasting Company had placed in operation eight broadcasting stations, partly connected by telephone wires, in various important cities. These stations were generally of higher power than most of those of the red network, and therefore with correspondingly greater coverage. They were formed into a second network completely connected by telephone lines, known as the blue network.

With the above as a nucleus, the National Broadcasting Company soon expanded rapidly. Stations were added, telephone line mileage increased, and the staff and studio facilities greatly enlarged, until today we have a completely capable technical and program organization which has to its credit a remarkable record of achievement. Its present status will now be taken up in detail below.

ORGANIZATION OF PERSONNEL

For the purpose of illustrating the interrelationship of the technical, commercial, and artistic personnel which forms the National Broadcasting Company, there is shown in Fig. 1 a general organization chart of the company.

The functions of some of the divisions are as follows:

(*1*) *Advisory Council.* In order that the public might cooperate in the development of the company, Owen D. Young invited eighteen recognized leaders in public life to act as an advisory council. This was done so that the organization might have the guidance of the ablest people in putting its facilities to the best possible use. This council also acts, when necessary, as a court of appeals in studying suggestions or complaints from the public. Special committees on agriculture, church activities, education, labor and women's activities have been formed from the membership of the council, which hold meetings at regular intervals and submit recommendations to the President. It is upon these recommendations that the program policies of the company are largely based.

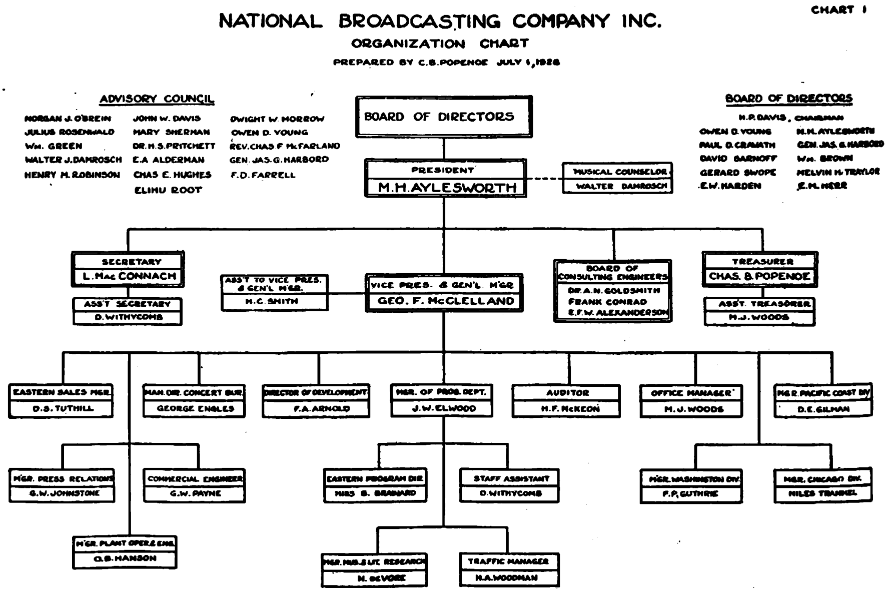


CHART I
NATIONAL BROADCASTING COMPANY INC.
ORGANIZATION CHART
PREPARED BY C.B.POPENOE JULY 1,1928
ADVISORY COUNCIL
MORGAN J. O'BREIN
JULIUS ROSENWALD
WM. GREEN
WALTER J.DAMROSCH
HENRY M.ROBINSON
JOHN W. DAVIS
MARY SHERMAN
DR.H.S.PRITCHETT
E.A ALDERMAN
CHAS E. HUGHES
ELIHU ROOT
DWIGHT W. MORROW
OWEN D. YOUNG
REV.CHAS F. McFARLAND
GEN. JAS.G.HARBORD
F.D.FARRELL
BOARD OF DIRECTORS
BOARD OF DIRECTORS
M.P.DAVIS, CHAIRMAN
OWEN D. YOUNG
PAUL D.CRAVATH
DAVID SARNOFF
GERARD SWOPE
E.W.HARDEN
M.H.AYLESWORTH
GEN. JAS. G. HARBORD
WM. BROWN
MELVIN H. TRAYLOR
E.M. HERR
PRESIDENT
M.H.AYLESWORTH
MUSICAL COUNSELOR
WALTER DAMROSCH
SECRETARY
L.MAC CONNACH
ASS'T SECRETARY
D.WITHYCOMB
ASS'T TO VICE PRES. & GEN'L MGR.
H.C.SMITH
VICE PRES. & GEN'L MGR
GEO. F. McCLELLAND
BOARD OF CONSULTING ENGINEERS
DR.A.N.GOLDSMITH
FRANK CONRAD
E.F.W. ALEXANDERSON
TREASURER
CHAS. B.POPENOE
ASS'T. TREASURER
M.J.WOODS
EASTERN SALES MGR.
D.S.TUTHILL
MAN.DIR. CONCERT BUR.
GEORGE ENGLES
DIRECTOR OF DEVELOPMENT
F.A.ARNOLD
MGR. OF PROG. DEPT.
J.W.ELWOOD
AUDITOR
H.F.McKEON
OFFICE MANAGER
M.J.WOODS
MGR.PACIFIC COAST DIV
D.E.GILMAN
MGR. PRESS RELATIONS
G.W.JOHNSTONE
COMMERCIAL ENGINEER
G.W.PAYNE
EASTERN PROGRAM DIR.
MISS B. BRAINARD
STAFF ASSISTANT
D.WITHYCOMB
MGR.WASHINGTON DIV.
F.P.GUTHRIE
MGR. CHICAGO DIV.
NILES TRAMMEL
MGR. PLANT OPER.& ENG.
O.B.HANSON
MGR.MUS.& LIT. RESEARCH
N. deVORE
TRAFFIC MANAGER
H.A.WOODMAN

Fig. 1

(*2*) *Program Department.* This is in itself an extensive organization, including trained musicians, continuity writers, announcers, vocal and instrumental artists and three complete "stock" companies—the National Grand Opera Company, the National Light Opera Company, and the National Players. There is a Department of Musical and Literary Research to investigate and authenticate the various selections and to guard against infringement of copyrights. There is an extensive library, which provides music of every description, and this department regularly composes special music for individual programs.

The program department, in addition to the basic duty of preparing the material and selecting the artists for the program, must also arrange and conduct as many rehearsals as are necessary to insure adequate presentation. Then, finally, before the program goes "on the air," it is picked up and reproduced on loud speakers in the monitoring rooms, to which the program director and his assistants listen attentively, so that the performance as it will be heard may realize the full intention of the program.

(3) *Plant Operation and Engineering.* This is the technical department of the company, and its functions include all the work connected with the installation and operation of the technical equipment, supervision of the quality of outgoing programs, observation of the programs delivered from the telephone wires connecting the various networks, and maintenance of telegraphic contact with all of the outlet stations connected to these networks. The entire department comprises 106 engineers, operators, maintenance men, draftsmen, machinists and similar personnel, located partly at the New York City headquarters, and in other cases in various cities of the United States where studio facilities are maintained or where important events may occur.

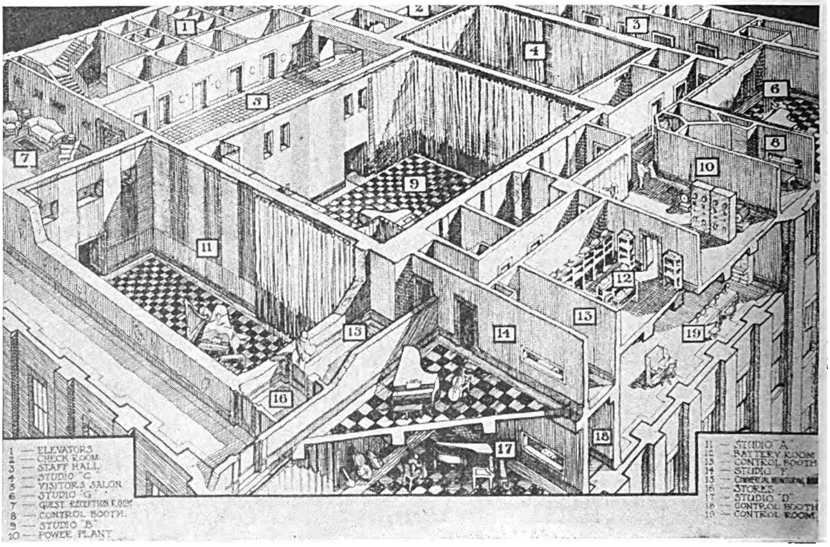
Fig. 2—National Broadcasting Company's building.

The functions of other departments are mainly self-evident from their titles.

It may be of interest in passing to mention the fact that the entire personnel of the company is now in excess of 600 persons.

STUDIO FACILITIES AND APPARATUS

The principal New York City studios of the National Broadcasting Company are situated at 711 Fifth Avenue, which is close to the city's musical, theatrical, and artistic centers. During 1927 the company arranged with the constructors of a new 16-story building at this location to have the four uppermost floors especially designed so as to house its broadcasting studios, offices, and engineering staffs. This building represents the first comprehensive, scientific effort to create an ideal broadcasting center, provided with every known facility for the advancement of the art. The design is due to Mr. O. B. Hanson, Manager of Plant Operation and Engineering.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1 — ELEVATORS
2 — CHECK ROOM
3 — STAFF HALL
4 — STUDIO "C"
5 — VISITORS SALON
6 — STUDIO "G"
7 — GUEST RECEPTION ROOM
8 — CONTROL BOOTH.
9 — STUDIO "B"
10 — POWER PLANT
11 — STUDIO "A"
12 — BATTERY ROOM
13 — CONTROL BOOTH
14 — STUDIO "F"
15 — COMMERCIAL MONITORING
16 — STORES
17 — STUDIO "D"
18 — CONTROL BOOTH
19 — CONTROL ROOM.

Fig. 3—New studios and control of the National Broadcasting Company at 711 Fifth Avenue, New York City.

The twelfth floor of this building houses the general offices of various departments, including the Program Department and library. The thirteenth and fourteenth floors are occupied by studios, reception rooms, control rooms, and the offices of the Plant Operations and Engineering Department. A general view of the building is shown in Fig. 2 and a cross-sectional view of these floors is shown in Fig. 3.

Studios *A*, *B*, and *C*, which are numbered in the diagram 11, 9, and 4, respectively, open directly on the main foyer of the thirteenth floor and are readily accessible from the reception room. The ceilings of these three studios are nineteen feet high extending through to the fourteenth floor. Observation windows which can be seen in the drawing in guest reception room (7) and the visitors' salon (5), on the fourteenth floor, permit visitors to see the artists while they are broadcasting. Two other studios, *D* (17) and *E*, which is not visible in the drawing, are on the thirteenth floor in addition to a main plant control room (19), foyer and artists' reception rooms. The area of this floor is approximately 13,000 square feet.

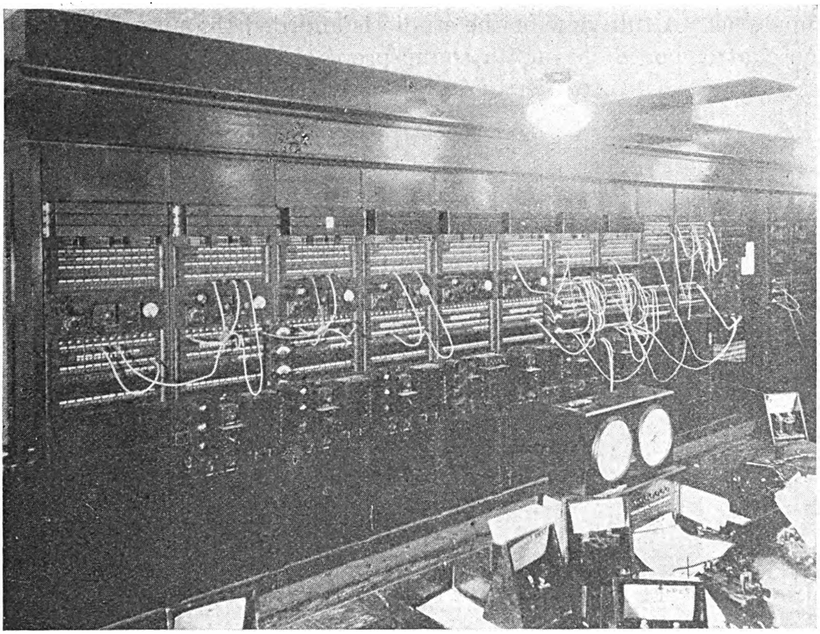


Fig. 4—Partial general view of main control room at National Broadcasting Company

Studio *B* is the largest on the thirteenth floor, measuring fifty-one feet long and thirty-six feet wide. The other studios, *A* and *C*, measure forty feet long by twenty-five feet wide. Studios *D* and *E* are located on the Fifty-fifth Street side of the building and are approached by wide corridors leading off the main foyer. These smaller studios are twenty-one feet by thirty-two, with ceilings eleven feet high.

The architectural treatment of the studios is distinctive and varied, the architect having striven to create a different effect in each of the studios. A different shade of color is used in each, studios *A*, *B*, and *C* being hung respectively in light fawn, old red and gold ochre drapes.

Studios *D* and *E*, the two smaller studios on this floor, are draped respectively in cadmium and burnt orange. With a lacquered cork tile floor the studios have the appearance of ballrooms. One or two small rugs are used in each studio.

Orchestral grouping has been simplified by placing numbers and letters along the mop board in all studios which enable the studio staff to designate positions by giving the performer a number and letter, rather than having to approximate his position to the microphone.

Each studio has a separate control room (13), (18), and (8), with apparatus. A full view of the studio is afforded the man on duty in the control booth through a sound-proof observation window. The announcer's controls are located adjacent to the booth so that he is in constant touch with the monitor. A master control room, which contains the bulk of the apparatus, is centrally located to the five studios enabling quick access to the individual control and monitoring rooms and studio.

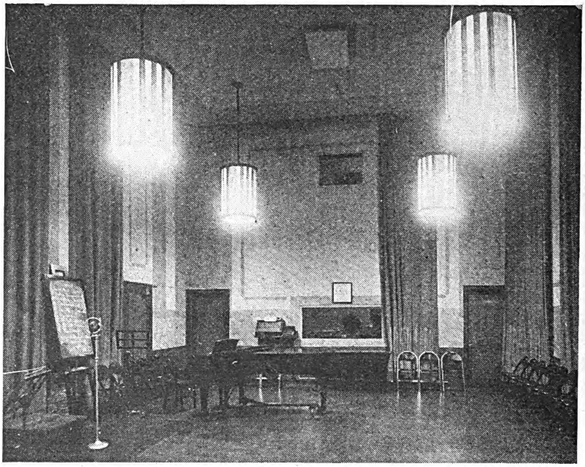


Fig. 5—General view of one of the small studios.

Amplifiers and detail apparatus of control are located in the central plant control room (19). This is where the wire networks terminate and where the various programs picked up throughout the metropolitan area are coordinated. The master control supervisor is on duty here at a desk resembling the console of a grand organ. Through an intricate system of switches and "jacks" he supervises the individual studio controls. Other equipment located here includes a temperature recorder for all studios, a master clock for synchronizing time schedules for the wire networks, and special testing apparatus required in broadcasting network programs. The ceiling of this room has been acoustically treated in a manner similar to that used in the studios in order that the supervisor can judge the quality of transmission.

On the fourteenth floor immediately above this room are the generators, master power boards, storage batteries, and the cross-connection room (12) and (10), where all the wires carrying either program material, or power, are distributed to the various parts of the building. Two more studios are on this floor, studios *F* (14) and *G* (16), which are the same size as *D* (17) and *E* (not shown) on the thirteenth floor. Due to the height of the ceilings of *A* (11), *B* (9), and *C* (4), there are no studios immediately over these three on the fourteenth floor. In addition to the foyer and visitors' reception rooms there are dressing rooms and shower baths for members of the NBC staff and also several green rooms for distinguished artists and internationally known celebrities.

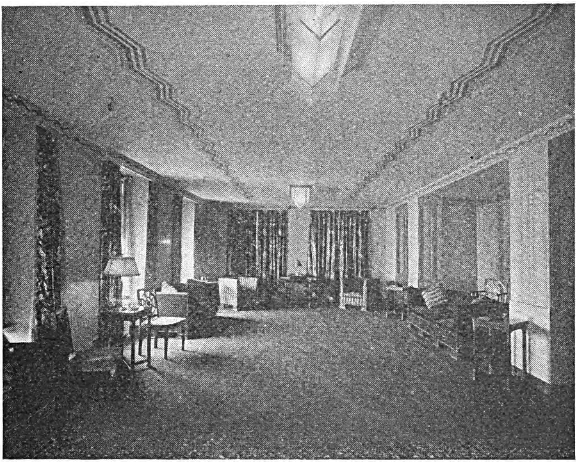


Fig. 6—One of the reception rooms.

On the fifteenth floor, which is not shown in the sketch, is located the largest studio, with dimensions of 40 by 80 ft., and with a 20-ft. ceiling. This is a small auditorium and is capable of seating approximately 250 people, should this be desired at any time. An indirect lighting system has been provided which utilizes more than 700 incandescent lamps which are used to create various color effects as desired. A stage has been provided at one end, and a pipe organ has been included in the construction of this studio. A spacious foyer opens into this auditorium-studio. The rest of this floor is used for offices of the National Broadcasting Company's executives.

The acoustical treatment of each studio consists of acoustical plaster on the walls and ceiling, the cork tile floor, and a small quantity of draperies and rugs. The latter are placed in back of and on the walls near the microphone, so that the artists work in a relatively "live" studio space, while the space in the immediate region of the microphone is more heavily damped. This method of operation is preferable to a uniformly highly damped studio, as it provides satisfactory acoustical conditions for both artists and microphone. The desirability of a "live" or moderately reflecting working space for performers and an acoustically "dead" space for listeners, in theatres or concert halls, has been recently brought out in the work of Professor F. R. Watson of the University of Illinois.[1] Similar arrangements have proved equally desirable in broadcasting studios, where the microphone takes the place of the listener.

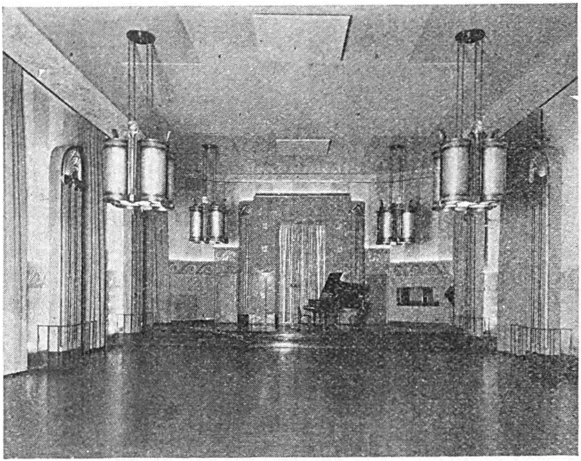


Fig. 7—General view of auditorium on 15th floor.

The placement of artists or orchestras with respect to the microphone in these studios has been thoroughly covered in an article[2] by Carl Dreher, staff engineer of the National Broadcasting Company, and will therefore not be gone into here.

A number of photographs are reproduced in Figs. 4, 5, 6, and 7,

[1] F. R. Watson, "Ideal Auditorium Acoustics," Jour. of the American Institute of Architects; July, 1928; or *Science*, p. 335, March 30, 1928.

[2] Carl Dreher, "Acoustics and Microphone Placing in Broadcast Studios," Proc. of the Radio Club of America, 5, numbers 5 and 6, May and June, 1928.

which show a portion of the main control room, some of the studios and a reception room.

The control room equipment consists of high quality amplifiers, volume indicators, and metering equipment of the standard broadcasting type, connecting to the individual small control rooms associated with the various studios. By means of this apparatus, supervision is maintained over the outgoing programs, but the actual regulation of intensity and placing of microphones is carried on in the small control rooms. The main control room also maintains contact, by means of telegraph circuits, with all of the radio stations associated with the various wire networks and supervises the linking up of the broadcasting telephone circuits which compose these networks. More detailed descriptions of this type of apparatus are given in the paper by D. G. Little, which appears as part of this symposium.

An interesting part of the plant is the "air conditioning" system. This was manufactured and installed by the Carrier Engineering Corporation of Newark, N. J. The equipment supplies fresh air at a constant temperature and humidity to all the studios and control rooms. The air taken in from out-of-doors is first washed free of dirt and dust; next it is dehumidified (that is, the excess of moisture over a certain amount is removed by cooling); and then the air is distributed at a temperature such that a comfortable room temperature is obtained. In summer the air is cooled to the proper temperature, and in winter it is heated by heating the cleansing water. In addition, it has recently been found advantageous to introduce ozone into the air, so as to precipitate smoke particles and to impart an exhilarating effect to the artists and staff.

WIRE FACILITIES AND OUTLET STATIONS

The programs originating in the New York studios, or in any other part of the United States, are supplied over a large telephone wire network (leased from the American Telephone and Telegraph Company) to numerous radio stations of various powers. Three radio stations are owned or leased by the National Broadcasting Company itself (WEAF and WJZ near New York, and WRC in Washington, D. C.), four are owned by the Westinghouse Electric and Manufacturing Company (KDKA in Pittsburgh, Pa., KYW in Chicago, Ill., WBZ in Springfield, Mass., and WBZA in Boston, Mass.), and three are owned by the General Electric Company (WGY in Schenectady, N. Y., KOA in Denver, Colo., and KGO in Oakland, Cal.). The remainder, now numbering 48 stations, are owned by a variety of interests including newspapers, automobile sales agencies, radio and electrical companies, municipalities, public utilities, department stores, and hotels.

The National Broadcasting Company maintains studios in Washington, Chicago, and San Francisco. In these cities permanent telephone wires are also installed to theatres or other places where program material may originate. In addition, temporary wire connections are set up to any cities in which an event of importance will occur, and connected into the station network.

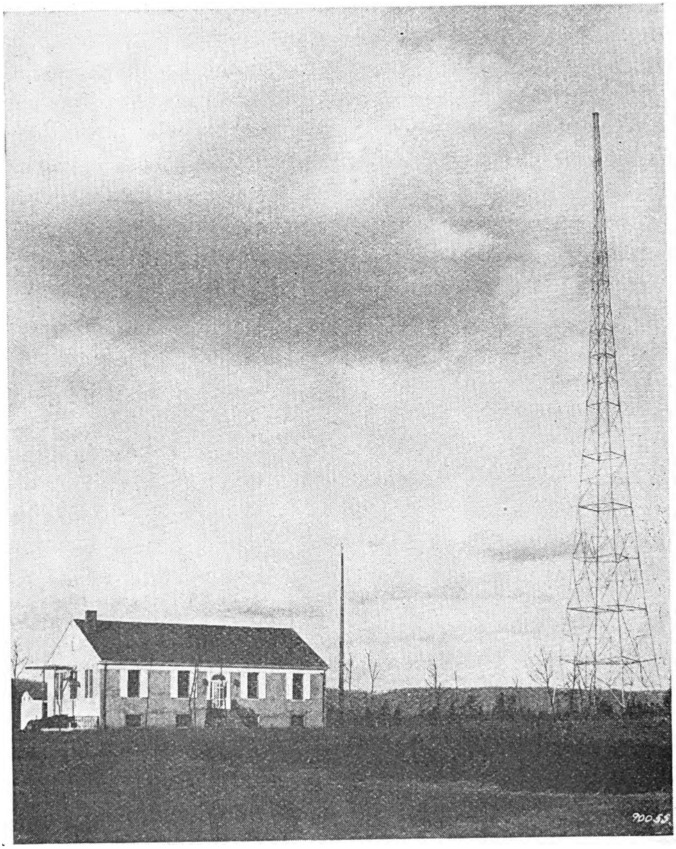
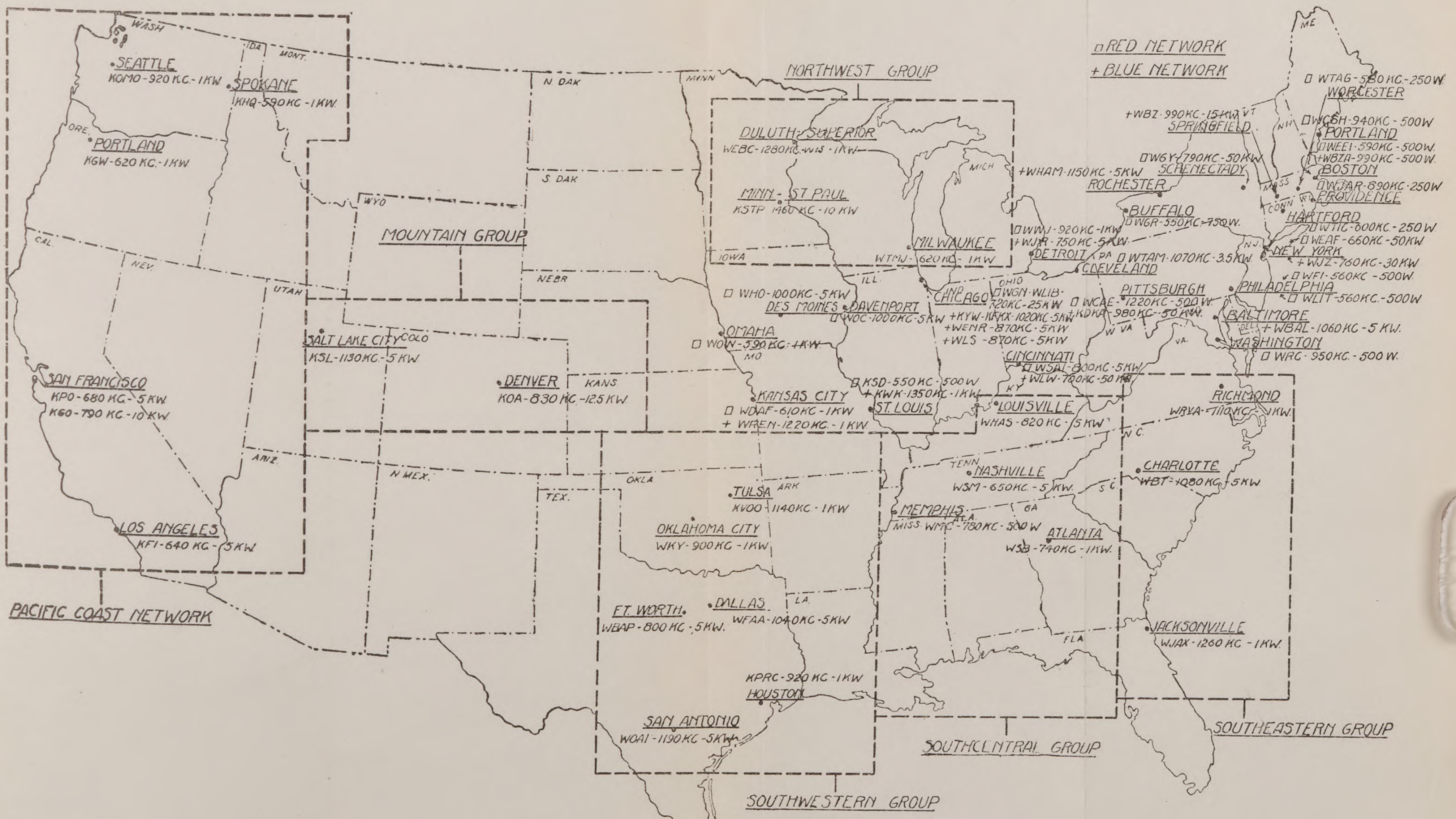


Fig. 9—General view of transmitter house and one of the antenna towers, Station WJZ.

Fig. 8 illustrates the locations of the various outlet stations.

Some facts in connection with this system may be of interest.


□ RED NETWORK
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PACIFIC COAST NETWORK
MOUNTAIN GROUP
NORTHWEST GROUP
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RI
SEATTLE
KOMO-920 KC.-1KW
SPOKANE
KHQ-590 KC-1KW.
PORTLAND
KGW-620 KC.-1KW
SAN FRANCISCO
KPO-680 KC.-5KW.
KGO-790 KC.-10KW
LOS ANGELES
KFI-640 KC.-5KW
SALT LAKE CITY
KSL-1130KC-5KW.
DENVER
KOA-830 KC.-12.5KW.
DULUTH-SUPERIOR
WEBC-1280KC-WIS-1KW
MINN.-ST PAUL
KSTP 1460 KC-10 KW
MILWAUKEE
WTMJ-620KC-1KW
□ WHO-1000KC-5KW
DES MOINES
DAVENPORT
□ WOC-1000KC-5KW
OMAHA
□ WOW-590 KC-1KW
KANSAS CITY
□ WDAF-610KC-1KW
+ WREN-1220KC.-1KW.
□ KSD-550KC-500W
+ KWK-1350KC-1KW
ST. LOUIS
CHICAGO
□ WGN-WLIB-720KC-25KW
+KYW-KFKX-1020KC-5KW
+WENR-870KC-5KW
+WLS-870KC-5KW
CINCINNATI
□ WSAI-800KC-5KW
+WLW-700KC-50KW
LOUISVILLE
WHAS-820KC-5KW
□ WWJ-920KC-1KW
+WJR-750KC-5KW.
DETROIT
□ WTAM-1070KC-3.5KW.
CLEVELAND
PITTSBURGH
□ WCAE-1220KC-500W.
+KDKA-980KC-50KW.
+WHAM-1150KC-5KW
ROCHESTER
BUFFALO
□ WGR-550KC-750W.
□ WGY-790KC-50KW.
SCHENECTADY
+WBZ-990KC-15KW
SPRINGFIELD
□ WTAG-580KC.-250W.
WORCESTER
□ WCSH-940KC-500W
PORTLAND
□ WEEI-590KC-500W.
+WBZA-990KC-500W.
BOSTON
□ WJAR-890KC-250W
PROVIDENCE
HARTFORD
□ WTIC-600KC.-250W
□ WEAF-660KC-50KW.
NEW YORK
+WJZ-760KC-30KW
□ WFI-560KC-500W.
PHILADELPHIA
□ WLIT-560KC.-500W
BALTIMORE
+WBAL-1060KC-5KW.
WASHINGTON
□ WRC-950KC.-500W.
RICHMOND
WRVA-1110KC-1KW.
CHARLOTTE
WBT-1080KC-5KW
NASHVILLE
WSM-650KC.-5KW.
MEMPHIS
WMC-780KC-500W
ATLANTA
WSB-740KC-1KW.
JACKSONVILLE
WJAX-1260 KC-1KW.
TULSA
KVOO 1140KC-1KW
OKLAHOMA CITY
WKY-900KC-1KW
FT. WORTH.
WBAP-800KC-5KW.
DALLAS
WFAA-1040KC-5KW
KPRC-920KC-1KW
HOUSTON
SAN ANTONIO
WOAI-1190KC-5KW

Fig. 8—Network stations of National Broadcasting Company

The total wire lines are in excess of 10,000 miles. The rental charges for these lines in 1927 were in excess of $1,350,000.

During 1927, also, the expenditure for radio programs to supply this network was about $6,000,000. Of this, over $2,000,000 was spent for talent alone on "sponsored" programs presented by some fifty American industries who are clients of the company.

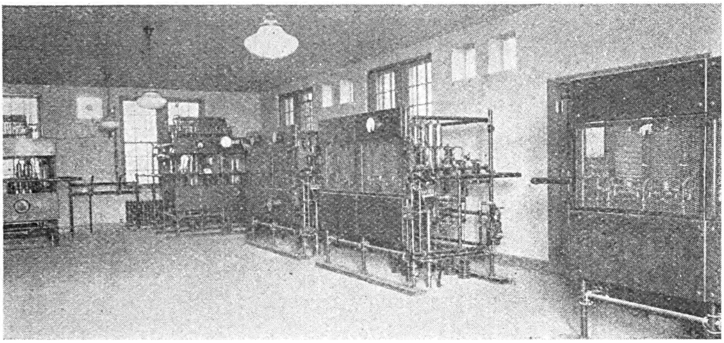

Fig. 10—General view of transmitting apparatus, station WJZ.

The programs supplied are either "sponsored" or "sustaining." The former are paid for by industrial establishments, for the purpose of creating goodwill towards their products. The latter are furnished by the National Broadcasting Company to occupy those periods during the day when sponsored programs are not available. The outlet stations share with the National Broadcasting Company in the payments received for the sponsored programs, but make moderate payments to the National Broadcasting Company for the sustaining programs. As it works out at the present time, the sponsored program payments to the stations more than compensate for their payments for

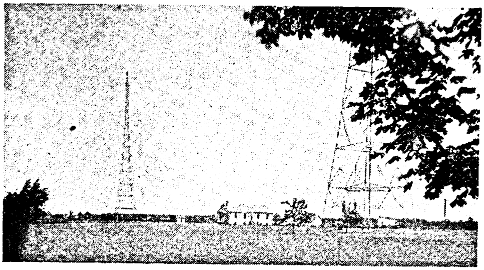

Fig. 11—General view of transmitter house and antenna towers at station WEAF.

sustaining programs, so that the system operates profitably, at least so far as the outlet stations are concerned. The National Broadcasting Company itself operated at a considerable deficit during the first year of its existence, but as sponsored programs have come to occupy a greater proportion of the total time, operations have become less unprofitable.

As exemplifying typical practice in high power American broadcasting stations, some views of the National Broadcasting Company's New York Stations, WJZ and WEAF, are shown in the accompanying figures.

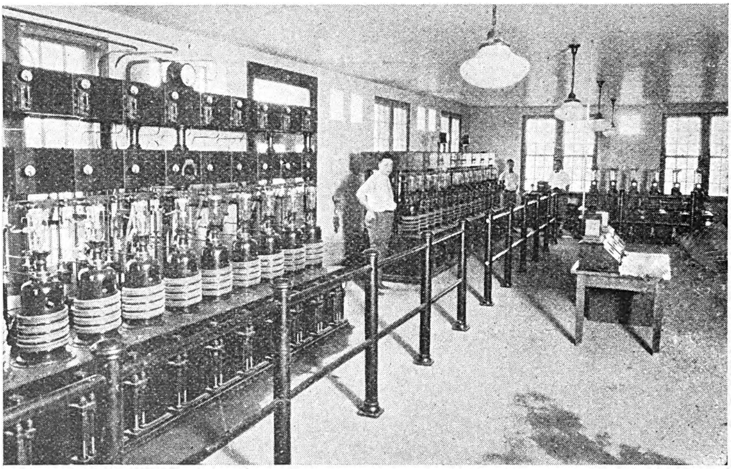

Fig. 12—Transmitting apparatus at station WEAF.

Station WJZ produces an antenna power of 30,000 watts and is located at Bound Brook, N.J., about 35 miles from New York City. Station WEAF, with an antenna power of 50,000 watts, is located at Bellmore, N. Y. (on Long Island), about 25 miles from New York City.

Fig. 9 is a general view of the transmitter building and of one of the antenna towers of WJZ. Fig. 10 shows the transmitting equipment. Power transformers, motor-generator sets, and other heavy apparatus are located in the basement of the building, immediately below the equipment shown in Fig. 10.

Fig. 11 is a general view of the station WEAF, and Fig. 12 shows the transmitting apparatus. As in the case of WJZ, the heavy equipment is located in the basement of the building, below the equipment shown.

TYPICAL PROGRAM ACHIEVEMENTS OF THE NATIONAL BROADCASTING COMPANY

Among the outstanding events in broadcasting made possible through the organization of the National Broadcasting Company were the following:

(1) The inaugural program of the company on November 15, 1926, was heard by an audience of more than 10,000,000 people. This program featured such great operatic stars as Titta Ruffo and Mary Garden, such theatrical celebrities as Weber and Fields, Will Rogers, Walter Damrosch and the New York Symphony Orchestra, Harold Bauer, the distinguished concert pianist, Edwin Franko Goldman and his band, as well as leading organizations in the field of more popular music. The event proved that there were no physical confines to the broadcasting studio. Mary Garden's voice was "picked up" from Chicago; Will Rogers spoke from Independence, Kansas, and the entire program was rendered as though hundreds of miles did not separate these performers from the broadcasting stations.

(2) On New Year's Day, January 1, 1927, there was broadcast for the first time in the history of the service an event originating on the Pacific coast, through the entire network. Millions of listeners on the eastern seaboard heard in the cold of midwinter a play-by-play report of the football game between Leland Stanford and Alabama Universities, which is played annually on New Year's Day in the Bowl of Roses at Pasadena, California.

(3) The first nationwide transmission of grand opera from the stage was accomplished by the National Broadcasting Company on January 21, 1927. This was the reproduction of the performance of the opera, *Faust*, presented in the Civic Auditorium in Chicago.

(4) On February 22, 1927, the address of the President of the United States, in commemoration of the anniversary of the birthday of George Washington, was broadcast through forty-two stations to an estimated audience of more than twenty millions of people.

(5) In June, 1927, during the Mississippi flood, the National Broadcasting Company, in a splendid effort to galvanize the relief activities of the nation at large, broadcast a special Red Cross relief program through its red and blue networks, reaching many millions of people and inspiring them to contribute their mites to the alleviation of the suffering. This event was notable in that it demonstrated strikingly the highly developed organization of the National Broadcasting Company and the entire feasibility of broadcasting programs of national interest from widely separated points in instantaneous sequence. At the beginning of this program an announcer in the studios

of WEAF at New York presented Secretary of Commerce Hoover. Instantly the skilled technical staff of the National Broadcasting Company shifted the circuits, and without pause Mr. Hoover, in the studios of WMC at Memphis, began to speak, from the very center of the flood area. After fifteen minutes the announcer in New York again went "on the air," introducing Al Jolson, the famous comedian, who began to speak from the studios of KYW in Chicago. Throughout this historic program there was not a single hitch, evidencing remarkably the highly developed capacities of the National Broadcasting Company, as well as the unimaginable potentialities of radio as an instrument of simultaneous nationwide intercommunication in time of need.

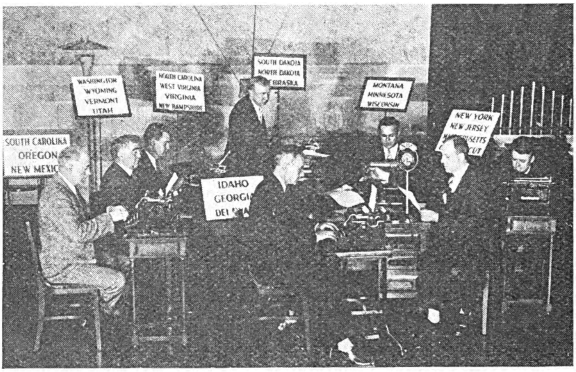
WASHINGTON
WYOMING
VERMONT
UTAH
NORTH CAROLINA
WEST VIRGINIA
VIRGINIA
NEW HAMPSHIRE
SOUTH DAKOTA
NORTH DAKOTA
NEBRASKA
MONTANA
MINNESOTA
WISCONSIN
NEW YORK
NEW JERSEY
SOUTH CAROLINA
OREGON
NEW MEXIC
IDAHO
GEORGI
DEL

Fig. 13—Broadcasting election returns throughout the United States.

(6) The historic airplane flights of Captain Charles A. Lindbergh and the other pioneers in transatlantic aviation were followed closely, and periodic bulletins broadcast to millions of waiting listeners as these flights progressed. The receptions accorded to these aviators on landing in the United States were similarly transmitted.

(7) On Thursday, July 21, 1927, through the courtesy of the Scripps-Howard newspapers, some forty millions of people in all parts of the world listened to Graham McNamee's graphic portrayal of the Dempsey-Sharkey bout at the Yankee Stadium, broadcast through the network of the National Broadcasting Company from fifty-one stations, two of which, WGY (Schenectady) and KDKA (Pittsburgh), broadcast on their short wavelengths as well as their regular frequencies, thus making the program available to interested auditors as remote as Europe, South America, Australia, and Africa.

(8) The Presidential election held during November, 1928, furnished a splendid opportunity for demonstrating the advantages

of the system of the National Broadcasting Company. The returns from the various states were received in a room at the headquarters of the National Broadcasting Company and reported to trained newspaper writers. These men maintained a constant flow of comment on the individual reports, which were handed to an announcer who read them into a microphone, whence they were transmitted through the entire network. Fig. 13 shows the group of men at work on that occasion.

Obviously, such immense coordinated undertakings would be entirely beyond the scope of any save a great organization like the National Broadcasting Company.

The "events" cited are outstanding because of their nationwide interest and the extraordinary facilities used to broadcast them; but the real influence of present day radio, as the National Broadcasting Company has developed it, lies in the less spectacular programs broadcast regularly through its own stations and varying numbers of the network stations. Among these are programs that at times include many prominent personages, such as the President of the United States, the Vice President, the Governors of states, the Prince of Wales, Premier Stanley Baldwin, eminent jurists, authors, journalists; famous operatic and concert stars; great conductors, symphony orchestras, and noted stars of the stage. Other types of programs include carefully adapted operatic and dramatic performances, general educational, and informative topics.

These programs, received regularly in millions of homes, are rationally and consistently extending and improving the culture of the nation. The effect is bound to be truly magnificent as the art progresses and evolves, since radio combines the functions of both amusement and education.

* * *

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SPEECH INPUT EQUIPMENT*

BY

D. G. LITTLE

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SPEECH input equipment is broadly understood to comprise the necessary apparatus to convert sound into electrical energy of a kind and amount suitable for use in broadcast transmitters. In addition to the equipment employed to pick up, amplify, and control the speech or program, supplementary apparatus for monitoring, intercommunicating, and power supply will also be described.

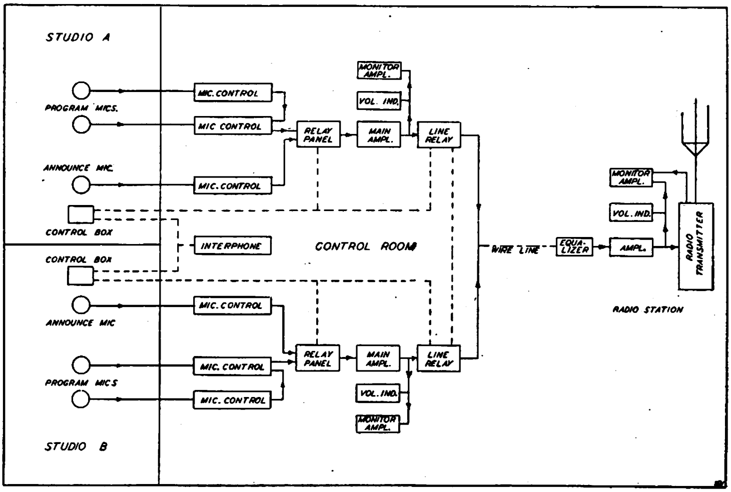
STUDIO A
PROGRAM MICS.
ANNOUNCE MIC
CONTROL BOX
CONTROL BOX
ANNOUNCE MIC
PROGRAM MICS
STUDIO B
MIC. CONTROL
MIC CONTROL
MIC. CONTROL
INTERPHONE
CONTROL ROOM
MIC. CONTROL
MIC. CONTROL
MIC. CONTROL
RELAY PANEL
MAIN AMPL.
LINE RELAY
MONITOR AMPL.
VOL. IND.
RELAY PANEL
MAIN AMPL.
LINE RELAY
VOL. IND.
MONITOR AMPL.
WIRE LINE
EQUALIZER
AMPL.
MONITOR AMPL.
VOL. IND.
RADIO TRANSMITTER
RADIO STATION

Fig. 1—Block diagram of studio and control room apparatus.

Each broadcast station has one or more studios where indoor programs are enacted. Each studio has its microphones connected to a group of amplifiers and control apparatus generally located in an adjacent control room. The program from any of these studios may be broadcast from several remotely located radio transmitters at the same time by the closing of interlocking relays which connect the various studio channels to the wire lines leading to the radio transmitting stations. The most elaborate and complete installation of this kind is that of the National Broadcasting Company at 711 Fifth Avenue, New York City, where eight studios are used to furnish programs to both the red and blue networks of wire lines connecting as many as sixty stations widely located throughout the United States. The detailed electrical features and characteristics of all the apparatus required for the operation of the National Broadcasting Company system in New York City are so involved as to be beyond the scope of this article. However, the general features of an ordinary broadcast system will be given in their relative order in the circuit from microphone to transmitter. This is best described by referring to the block diagram, Fig. 1.

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* Dewey decimal classification: R610. Original manuscript received by the Institute, April 11, 1929.

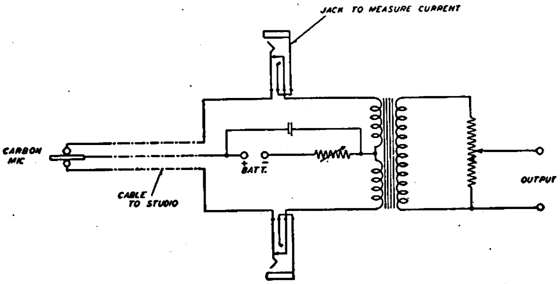

Fig. 2—Carbon microphone and control panel circuit diagram.

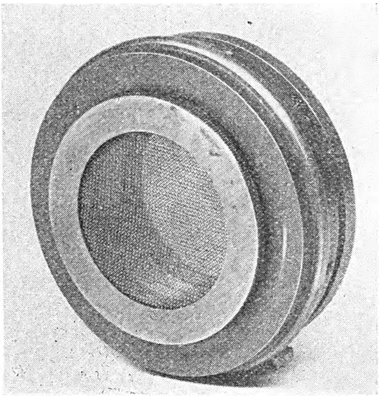

Fig. 3—RCA condenser microphone unit.

In general, frequencies between fifteen cycles and twenty thousand cycles per second are called audio frequencies, since sounds heard by the human ear are within this range. It is not necessary to cover this entire range of frequency in order to give quite satisfactory speech and music quality to our broadcast programs. Speech input equipment is usually designed to have a flat or uniform frequency characteristic from 60 to 6000 cycles. Performance characteristics of speech input equipment are usually given in TU (transmission units) which is a mathematical relation of the power output of a given amplifier or network to its power input.[1] Energy level, expressed in TU, refers to signal intensity or volume at a given point and time and is based upon an adopted standard of zero level in TU as being a power of 10 mw in six hundred ohms.

Microphones in use are of two general types: (a) carbon granule type, and (b) condenser type. Both are designed for high quality, that is, to have a substantially flat frequency characteristic from 60 to between 5000 and 10,000 cycles.

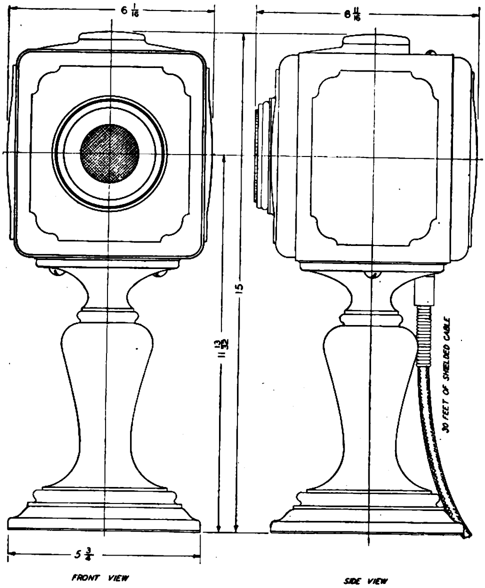

Fig. 4—Microphone amplifier and stand for table or desk mounting.

The carbon microphone employs two carbon variable resistance elements, called buttons, one on either side of a tightly stretched hard aluminum diaphragm. The circuit is shown in Fig. 2. In practice, the direct current through the carbon microphone flows from the stretched diaphragm to the buttons and is regulated by a rheostat on a microphone control panel to between 10 and 20 ma per button which assures the desired life of the device. The output level of this type of micro-

---

[1] Note: The gain or loss of energy in transmission units (TU) is expressed as follows:

$$\text{TU} = 10 \log_{10} P_1/P_2$$

Where P_1 = power at one set of terminals of the apparatus, and P_2 = power at at the other set of terminals.

phone is between −40 and −60 TU, depending upon the nature of the sound source and the distance from the microphone.

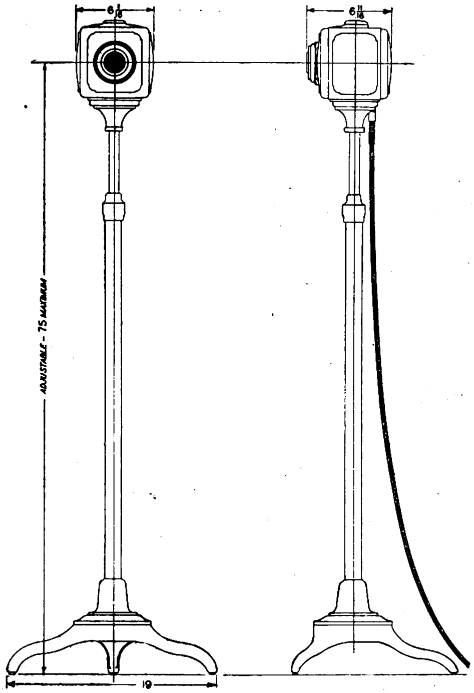
6 1/8
6 11/16
ADJUSTABLE - 75 MAXIMUM
19

Fig. 5—Microphone amplifier and stand.

The condenser microphone is fast replacing the carbon for program work, being desirable where the slightly greater ground tone of

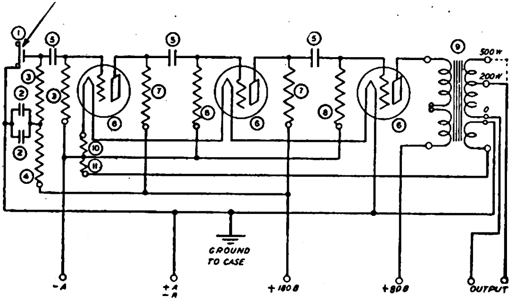
CONDENSER MICROPHONE
①
②
②
③
③
④
⑤
⑤
⑤
⑥
⑥
⑥
⑦
⑦
⑧
⑧
⑨
⑩
⑪
500 W
200 W
0
GROUND TO CASE
-A
+A
-A
+180 B
+80 B
OUTPUT

Fig. 6—Condenser microphone and amplifier circuit diagram.

the carbon type might be objectionable. A stretched diaphragm similar to that used in the carbon microphone is spaced about two thousandths of an inch from a carefully insulated back-plate, thus form-

which is amplified by a local amplifier to an energy level approximately that of the carbon microphone. Because of the very low energy level at the output of the condenser element, and the relatively low capacity of this element, it is common practice to locate the first or local amplifier close to or in the same unit with the condenser element. Fig. 3 shows the RCA type of condenser element; Figs. 4 and 5 show the amplifier in which this condenser element is mounted. Fig. 6 is the

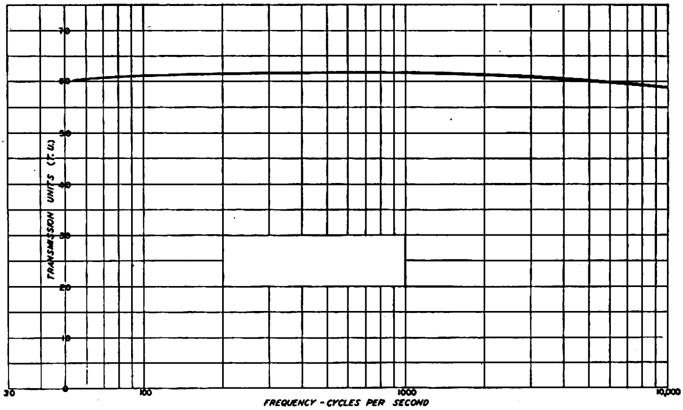

Fig. 8—12A, 3-stage amplifier frequency characteristic.

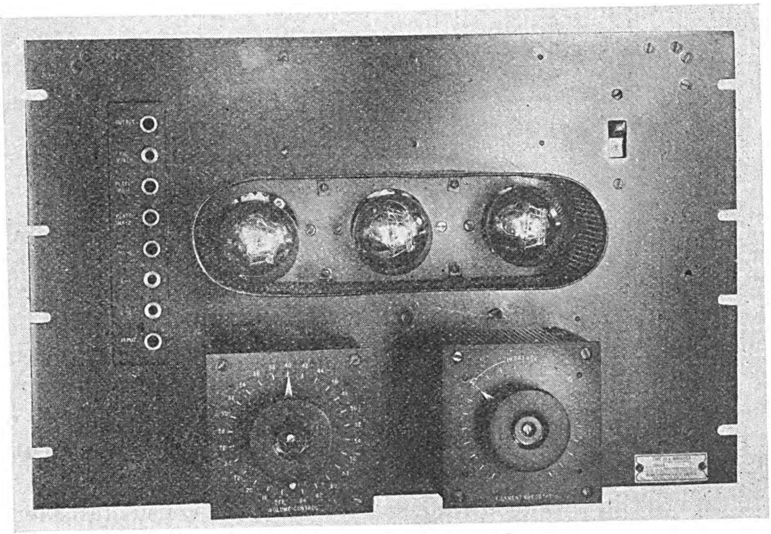

Fig. 9—12A, 3-stage amplifier.

circuit diagram. A five-conductor cable connects the complete microphone unit with its control panel where the amplifier tube filament voltage is regulated and the audio output adjusted.

The microphones described above are, of course, located in the studio or at the point of "pick-up." The outputs of the microphones are controlled, switched, and further amplified in the control room in panel type equipment mounted on vertical racks. Fig. 7 shows control room apparatus for two studios, in which a single set of amplifying and monitoring equipment is used for the outputs of both studios. The audio output of any microphone is adjustable by a potentiometer on its control panel. This adjustable output is connected through a relay panel and a jack panel to a multistage amplifier which raises the energy level to a value suitable for application to the telephone line going to the radio transmitter. This output energy is usually about zero level. The frequency characteristic of a typical three-stage amplifier is given in Fig. 8, and photograph of the unit in Fig. 9. The circuit diagram is shown in Fig. 10. The maximum gain from input to output terminals is between 60 and 64 TU, and the maximum output without overloading is a level of +16 TU.

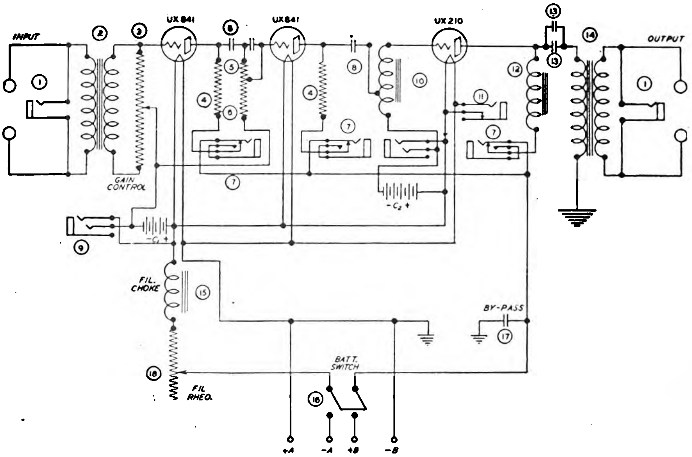

Fig. 10—21A, 3-stage amplifier circuit diagram.

In case the line between studio and station is of considerable length or is lead cable, a line compensating unit or filter must be used. This unit is usually known as a line equalizer and compensates for the frequency characteristic of the line which attenuates high frequencies more than low frequencies. A unit of this kind is shown in Fig. 11.

At the radio transmitting station, further amplification of the program is sometimes necessary in order to compensate for line losses and equalization losses and to raise the energy to approximately zero TU level, the input required for most transmitters.

Having followed the path of the program from the microphone in the studio through its various controls and amplification circuits to the radio transmitter, accessory, monitoring, and inter-communicating apparatus will now be considered. This accessory apparatus, although not traversed by the program itself, is of great importance. Fig. 1 shows this accessory apparatus as associated with the main audio channels.

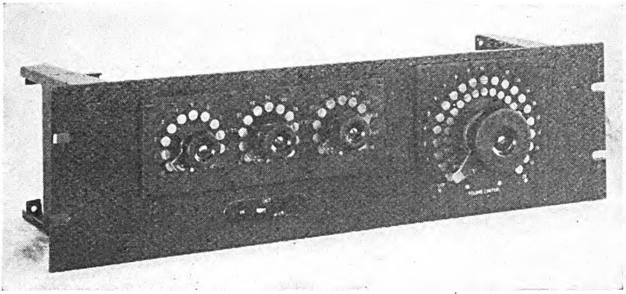

Fig. 11—Line compensating unit.

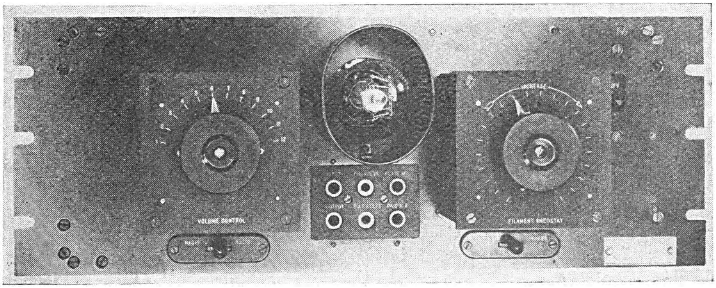

Fig. 12—Monitoring amplifier.

For conveying general information, such as the arrangements for the various parts of a program, etc., intercommunicating telephones connect the control room with the studios and with the radio station. A Morse telegraph circuit generally connects the control room and radio station. If the distance is short, this circuit may be "phantomed" onto the program line. The necessary switching of circuits during a program is usually made by the control room operator and the announcer through interlocked relays. These relays operate lamp signals between studios and control room and continuously display the circuit setup.

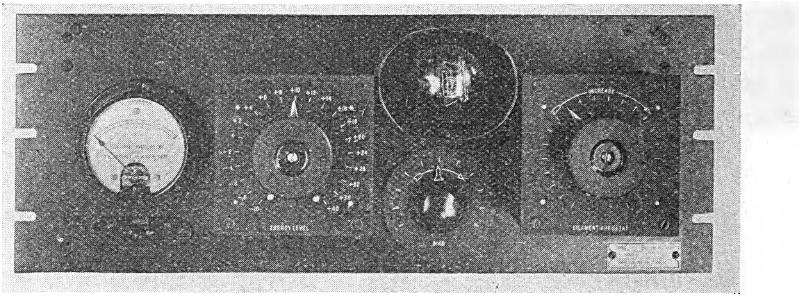

Fig. 13—Volume indicator.

In the control room, a monitoring amplifier and loud speaker keep the operator in touch with the program. Sometimes a second monitoring amplifier supplies loud speakers in the idle studios for auditions and permits an announcer of a following program to hear when the

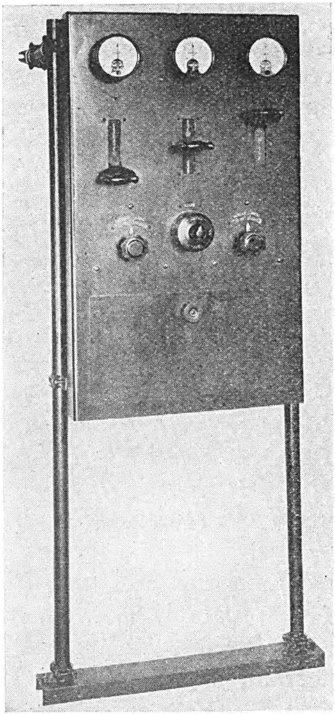

Fig. 14—29A battery charging panel.

preceding program ends. Otherwise, he depends upon lamp signals for this indication. In case of programs from two or more studios going simultaneously through the same control room, head telephones may be employed for monitoring. Fig. 12 shows one type of monitoring amplifier.

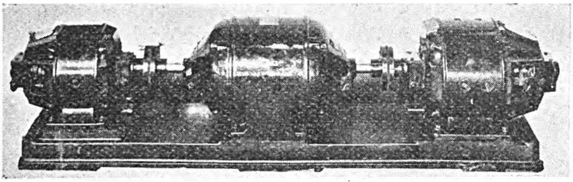

Fig. 15—Battery charging motor-generator set.

A volume indicator or vacuum-tube voltmeter calibrated in TU is shown in Fig. 13. At least one of these units is used at every studio control room and one at the station control room. At the studio control room, the volume indicator measures the level in TU delivered to the line; and at the station, it measures the level delivered to the transmitter. These units give visual indication of the signal level and allow the operators to adjust the "gains" of the amplifiers to the desired amount.

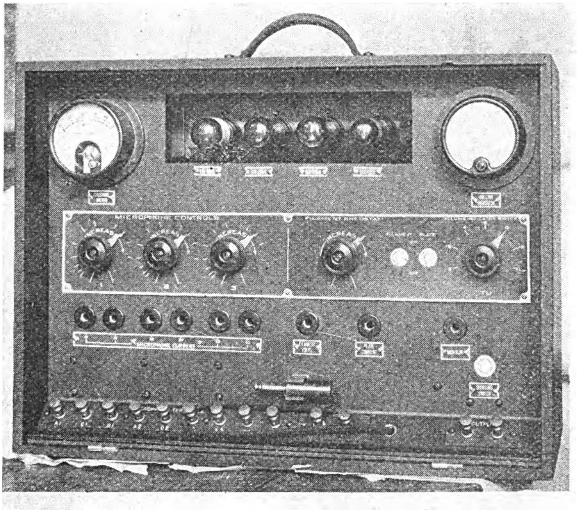

Fig. 16—Portable pickup amplifier, front.

Power for the vacuum tubes, used in the amplifiers, for the signal lights, relays, and other speech input apparatus, is obtained from storage batteries. A potential of twelve volts is used for the tube filaments, relays, and signal circuits, and 400 volts for the tube plate circuits. A three-unit motor-generator set with low- and high-voltage generators, and a battery charging panel are provided for the recharging of the batteries. The usual practice is to install two complete 12-volt batteries so that one may be charged while the other is in use. Battery installations are generally of a size sufficient to give twenty-four hours of service without recharging. Fig. 14 shows one type battery charging panel, and Fig. 15 shows its associated motor-generator set.

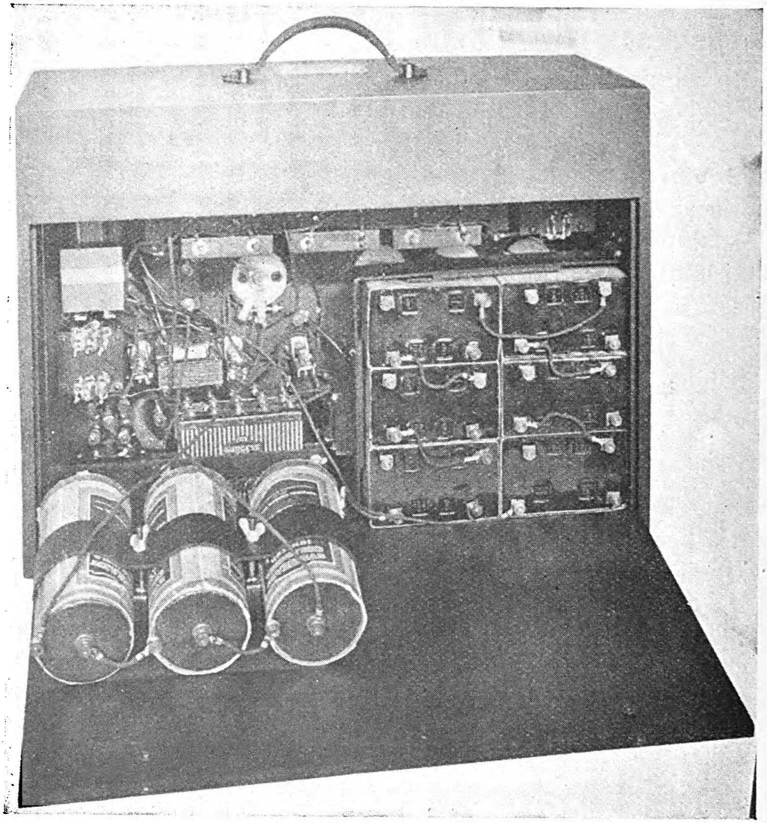

Fig. 17—Portable pickup amplifier, rear.

A portable type of combined amplifier and control unit has been developed for use at "outside" pickup points, such as, baseball and football games, celebrations, etc., also at churches, conventions, and in fact anywhere outside of a regularly equipped studio. One, two, or three carbon microphones are employed, or condenser microphones may be used by adding a battery box. Apparatus of this kind is transported to the place of pickup and connected to lines running to the studio control room or radio station, a short time before the program is scheduled to start. The apparatus is usually removed immediately following the program. See Figs. 16 and 17.

It is obvious that a radio receiver having a perfect fidelity characteristic and receiving from a transmitter with perfect characteristics will not reproduce a program satisfactorily unless this program is properly "picked up," amplified, and controlled by the speech input microphone and its associated equipment. Because of the fact that one radio station is heard by thousands of listeners and also that one microphone may be picking up a program for as many as sixty stations, the importance of correct design of speech input equipment is readily appreciated. In the design of speech input equipment, therefore, the cost is of secondary importance, and the very best of materials are employed. As an additional safeguard against trouble, the main equipment, amplifiers, and monitoring devices generally have their inputs and outputs normalled through jack panels so that in case of failure the defective panel may be electrically disconnected and a spare or duplicate panel put in service by merely inserting the proper patch cords between jacks. The success of the present broadcasting service in the United States is due in a large measure to the high quality of speech input energy delivered to the input terminals of the nation's broadcast transmitters and to the precautions taken to insure against interruptions to programs.

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WIRE LINE SYSTEMS FOR NATIONAL BROADCASTING*

By

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WHAT we are here considering, as an important factor in promoting national solidarity, is the tying together of a whole nation so that a single broadcast will instantly reach even the most remote points. Radio broadcasting stations (employing the more generally used frequencies) are essentially local distribution centers serving effectively points up to 50 miles (80 km), or, in favorable cases, 100 miles (160 km) or more from the radio transmitter. For the larger nations it is evidently necessary to make division into areas, locating a radio transmitter in each area for its coverage, and then to provide a network of circuits connecting the transmitters in the various areas with the point at which the broadcast originates. At the present time, wire telephone systems are employed almost exclusively for this national distribution of broadcasts. It is the purpose of this paper to discuss the wire networks which are now being provided in the United States by the Bell Telephone System.

In the United States at the present time (January 15, 1929) programs are being regularly distributed over extensive wire networks or "chains," as indicated on the map of Fig. 1. The various chains are usually referred to by colors. As a regular procedure most of these chains operate about six hours each day. Following are the numbers of radio stations served by each chain together with the lengths of telephone circuit involved. (An additional chain which operates only one hour each week is not included.)

	Radio Stations	Telephone Circuit Miles	
Red network	41	10,300	(16,600 km)
Purple network	41	8,450	(13,600 ")
Blue network	12	3,650	(5,900 ")
Green network	8	3,600	(5,800 ")
Orange network	5	1,700	(2,700 ")
Brown network	3	450	(700 ")
Total	110	28,150	(45,300 ")

On occasions when events of particular importance take place, several of the regular chains may be merged together and additional circuits added so as to pick up programs from various parts of the

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* Dewey decimal classification: R450. Original manuscript received by the Institute, January 31, 1929.

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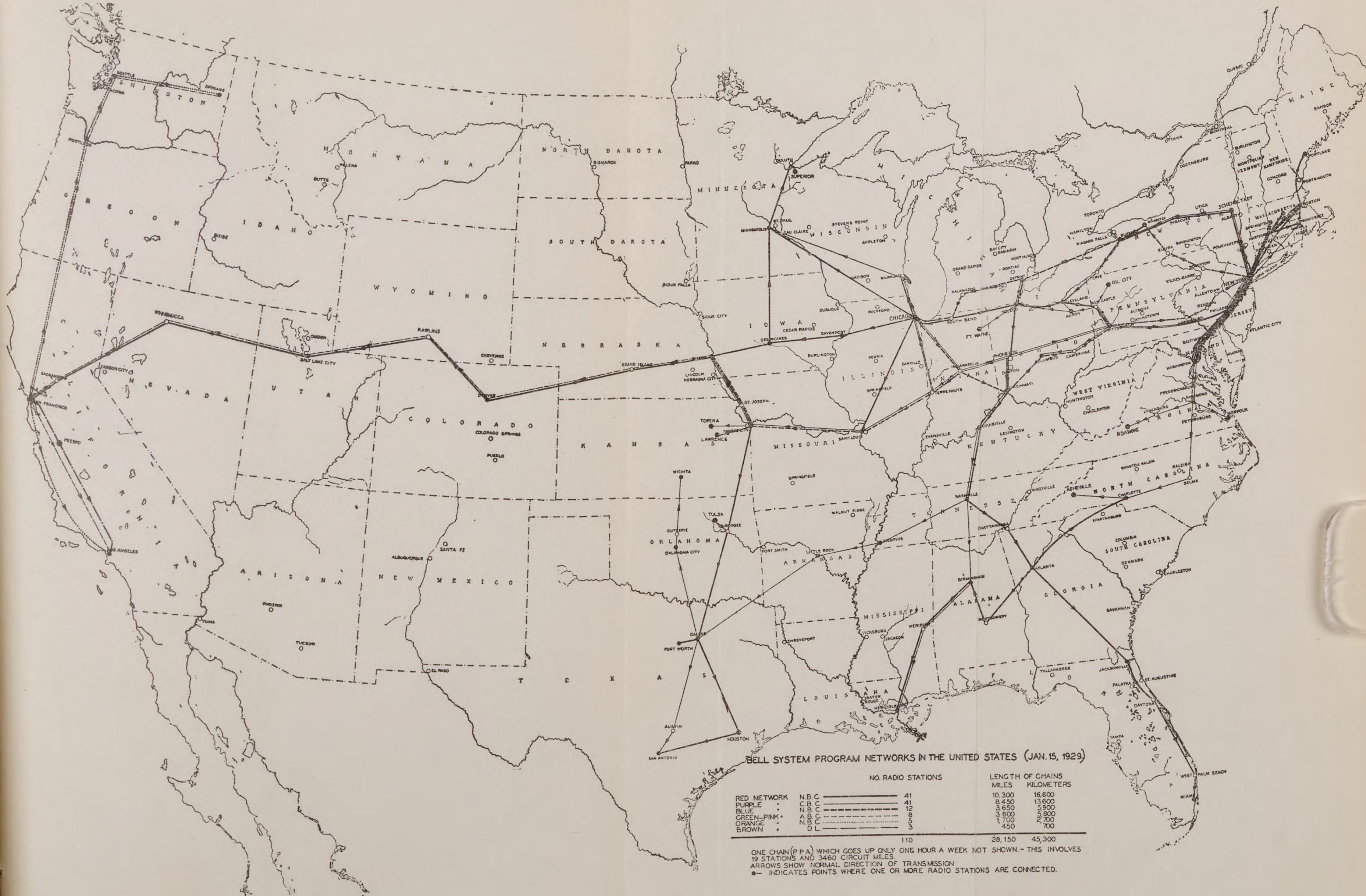


WASHINGTON
SEATTLE
TACOMA
SPOKANE
PORTLAND
OREGON
IDAHO
BOISE
MONTANA
HELENA
BUTTE
WYOMING
CHEYENNE
RAWLINS
NORTH DAKOTA
BISMARCK
FARGO
SOUTH DAKOTA
SIOUX FALLS
MINNESOTA
DULUTH
SUPERIOR
MINNEAPOLIS
ST. PAUL
EAU CLAIRE
WISCONSIN
STEVENS POINT
APPLETON
MADISON
MILWAUKEE
MICHIGAN
GRAND RAPIDS
BAY CITY
SAGINAW
PORT HURON
PONTIAC
DETROIT
KALAMAZOO
SOUTH BEND
FT. WAYNE
TOLEDO
CALIFORNIA
SACRAMENTO
SAN FRANCISCO
FRESNO
LOS ANGELES
NEVADA
WINNEMUCCA
CARSON CITY
UTAH
OGDEN
SALT LAKE CITY
COLORADO
DENVER
COLORADO SPRINGS
PUEBLO
NEBRASKA
GRAND ISLAND
LINCOLN
NEBRASKA CITY
OMAHA
IOWA
SIOUX CITY
DES MOINES
CEDAR RAPIDS
DAVENPORT
DUBUQUE
BURLINGTON
ROCKFORD
CHICAGO
PEORIA
DANVILLE
SPRINGFIELD
ILLINOIS
INDIANA
INDIANAPOLIS
TERRE HAUTE
KANSAS
TOPEKA
LAWRENCE
KANSAS CITY
ST. JOSEPH
WICHITA
MISSOURI
SAINT LOUIS
SPRINGFIELD
OKLAHOMA
TULSA
MUSKOGEE
GUTHRIE
OKLAHOMA CITY
ARKANSAS
FORT SMITH
LITTLE ROCK
WALNUT RIDGE
ARIZONA
PHOENIX
TUCSON
YUMA
NEW MEXICO
SANTA FE
ALBUQUERQUE
EL PASO
TEXAS
DALLAS
FORT WORTH
AUSTIN
SAN ANTONIO
HOUSTON
LOUISIANA
SHREVEPORT
BATON ROUGE
NEW ORLEANS
MISSISSIPPI
VICKSBURG
JACKSON
MERIDIAN
ALABAMA
BIRMINGHAM
MONTGOMERY
TENNESSEE
MEMPHIS
NASHVILLE
CHATTANOOGA
KNOXVILLE
KENTUCKY
LOUISVILLE
LEXINGTON
EVANSVILLE
OHIO
COLUMBUS
DAYTON
CINCINNATI
CLEVELAND
CAMBRIDGE
AKRON
WEST VIRGINIA
HUNTINGTON
CHARLESTON
PENNSYLVANIA
OIL CITY
ERIE
NEW CASTLE
PITTSBURGH
ALTOONA
JOHNSTOWN
HARRISBURG
WILKES-BARRE
ALLENTOWN
READING
PHILADELPHIA
NEW JERSEY
ATLANTIC CITY
BALTIMORE
WASHINGTON
VIRGINIA
RICHMOND
LYNCHBURG
ROANOKE
PETERSBURG
NORFOLK
FREDERICKSBURG
NORTH CAROLINA
WINSTON SALEM
RALEIGH
SELMA
ASHEVILLE
CHARLOTTE
SOUTH CAROLINA
SPARTANBURG
COLUMBIA
DENMARK
CHARLESTON
GEORGIA
ATLANTA
SAVANNAH
FLORIDA
TALLAHASSEE
JACKSONVILLE
ST. AUGUSTINE
PALATKA
DAYTONA
TAMPA
WEST PALM BEACH
MIAMI
NEW YORK
TORONTO
HAMILTON
NIAGARA FALLS
BUFFALO
ROCHESTER
SYRACUSE
UTICA
SCHENECTADY
ALBANY
ELMIRA
BINGHAMTON
POUGHKEEPSIE
OGDENSBURG
OTTAWA
MONTREAL
QUEBEC
VERMONT
BURLINGTON
MONTPELIER
NEW HAMPSHIRE
CONCORD
MAINE
BANGOR
PORTLAND
PORTSMOUTH
MASSACHUSETTS
BOSTON
SPRINGFIELD
WORCESTER
PROVIDENCE
HARTFORD
CONNECTICUT

BELL SYSTEM PROGRAM NETWORKS IN THE UNITED STATES (JAN. 15, 1929)

			NO. RADIO STATIONS	LENGTH OF CHAINS MILES	KILOMETERS
RED NETWORK	N.B.C.	━━━━━	41	10,300	16,600
PURPLE "	C.B.C.	———	41	8,450	13,600
BLUE "	N.B.C.	▬ ▬ ▬	12	3,650	5,900
GREEN–PINK "	A.B.C.	- - - -	8	3,600	5,800
ORANGE "	N.B.C.	·······	5	1,700	2,700
BROWN "	D.L.	—·—·—	3	450	700
			110	28,150	45,300

ONE CHAIN (P.P.A.) WHICH GOES UP ONLY ONE HOUR A WEEK NOT SHOWN.– THIS INVOLVES 19 STATIONS AND 3460 CIRCUIT MILES.
ARROWS SHOW NORMAL DIRECTION OF TRANSMISSION
●— INDICATES POINTS WHERE ONE OR MORE RADIO STATIONS ARE CONNECTED.

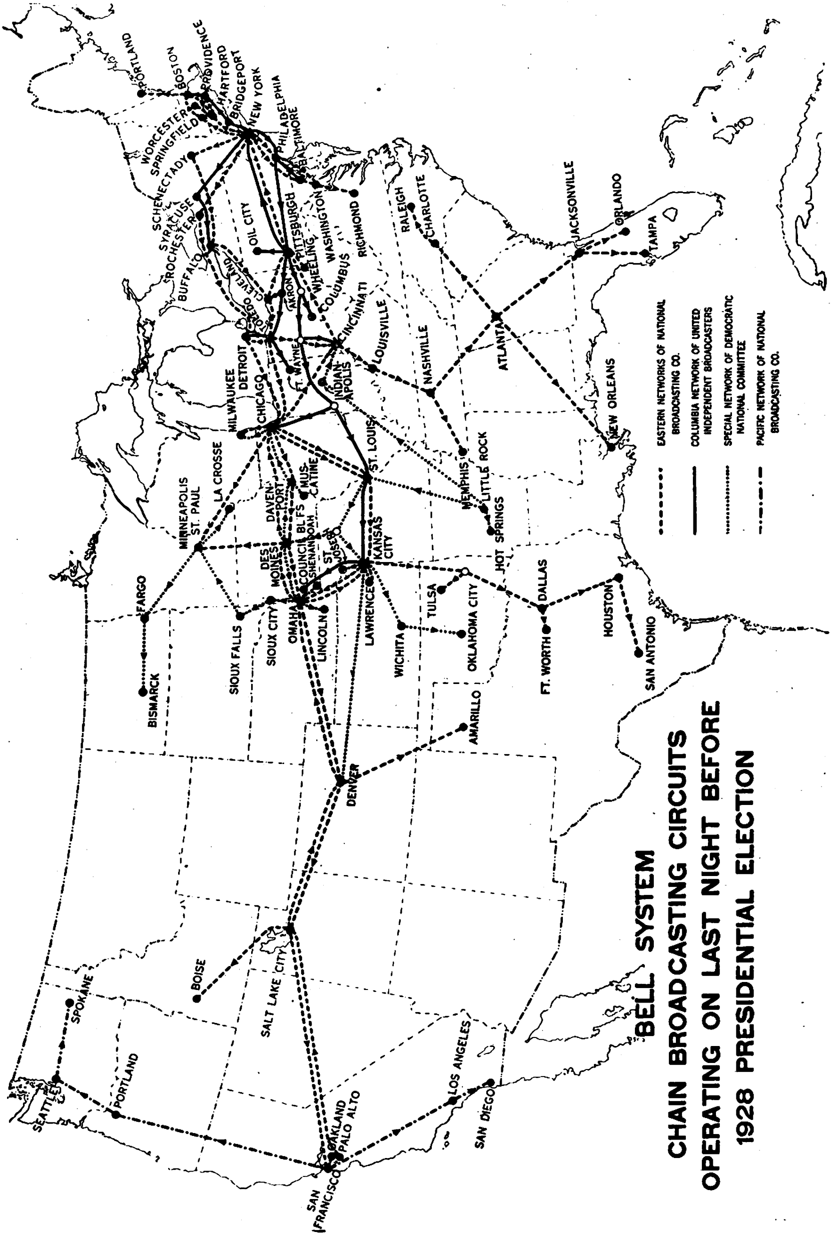
Fig. 1

country. For example, on November 5, 1928, the evening before the United States presidential election, the networks shown in Fig. 2 were in operation, about 85 radio stations being included. At various times during this evening, five separate programs were broadcast from several different points in New York City; Palo Alto, California; Little Rock, Arkansas; and Pittsburgh, Pennsylvania. The United States was thus virtually one great auditorium, with listeners estimated as no less than fifty million.

From the technical standpoint, program transmission circuits are, of course, very different from message telephone circuits. In the first place, message telephone circuits must be arranged so that to and fro conversations can take place practically instantaneously. Program transmission circuits on the contrary are single-direction transmission circuits. They are, therefore, not complicated by problems of electrical echo, singing and the like, which are ever present with long message telephone circuits. However, although free from the problems of two-way working, the design and operation problems of program transmission circuits are by no means easy as compared with those of message telephone circuits. On the contrary, in many respects, these problems are considerably more difficult, the reason being that the requirement as to approach to absolute fidelity of reproduction is much more severe than for message telephone circuits.

A frequency band width of 2,500 cycles furnishes, if properly utilized, a telephone circuit over which speech is transmitted very clearly so that conversations may be easily carried on. This band is not adequate, however, for program transmission because of the different character of the transmitted material. The bulk of present-day broadcast programs consists of musical selections, including a fair amount of high-grade material. To reproduce music, and particularly high-grade music, in a pleasing manner calls for a materially widened band. This wider band also gives a high degree of naturalness to speech which is particularly desirable when loud speakers are used for reception.

At the present time in the United States the frequency band which is transmitted over the long distance program chains extends from about 100 cycles to about 5,000 cycles. It is, of course, possible to transmit an even wider band than this, although the cost of the circuits will, of course, increase as the band is widened. In considering how wide the band should be, the complete system, including pickup apparatus, wire transmission line, radio transmitters, radio transmission paths through the ether, radio receiving apparatus and loud speakers must be considered. It seems probable that as the art progresses a band wider than the above will be found desirable. On the wire line



BELL SYSTEM
CHAIN BROADCASTING CIRCUITS
OPERATING ON LAST NIGHT BEFORE
1928 PRESIDENTIAL ELECTION

EASTERN NETWORKS OF NATIONAL BROADCASTING CO.
COLUMBIA NETWORK OF UNITED INDEPENDENT BROADCASTERS
SPECIAL NETWORK OF DEMOCRATIC NATIONAL COMMITTEE
PACIFIC NETWORK OF NATIONAL BROADCASTING CO.
PORTLAND
BOSTON
PROVIDENCE
HARTFORD
BRIDGEPORT
NEW YORK
WORCESTER
SPRINGFIELD
PHILADELPHIA
BALTIMORE
SCHENECTADY
SYRACUSE
ROCHESTER
BUFFALO
OIL CITY
PITTSBURGH
WASHINGTON
RICHMOND
RALEIGH
CHARLOTTE
JACKSONVILLE
ORLANDO
TAMPA
CLEVELAND
TOLEDO
AKRON
WHEELING
COLUMBUS
CINCINNATI
DETROIT
FT. WAYNE
LOUISVILLE
NASHVILLE
ATLANTA
MILWAUKEE
CHICAGO
INDIANAPOLIS
ST. LOUIS
NEW ORLEANS
MEMPHIS
LITTLE ROCK
HOT SPRINGS
LA CROSSE
MINNEAPOLIS ST. PAUL
DAVENPORT
MUSCATINE
DES MOINES
COUNCIL BLFS
SHENANDOAH
ST. JOSEPH
KANSAS CITY
FARGO
BISMARCK
SIOUX FALLS
SIOUX CITY
OMAHA
LINCOLN
LAWRENCE
WICHITA
TULSA
OKLAHOMA CITY
DALLAS
FT. WORTH
HOUSTON
SAN ANTONIO
AMARILLO
DENVER
SALT LAKE CITY
BOISE
SPOKANE
SEATTLE
PORTLAND
OAKLAND
PALO ALTO
SAN FRANCISCO
LOS ANGELES
SAN DIEGO

Fig. 2

systems, development work is going forward looking toward the possibility that such wider bands may be found desirable in the future. At the lower frequencies, where most people consider that improvement is particularly desirable, consideration is being given to the possible extension of the band down to 50 cycles and possibly lower. Consideration is also being given to the possible addition of two or three thousand cycles to the top of the band.

In addition to this broad band transmission requirement, program transmission circuits must be designed to handle wide ranges of volume, particularly for the transmission of musical programs. Much of the enjoyment in listening to good music appears to come from the ranges of volume, so that in order to deliver such musical programs properly these ranges of volume must be preserved in large part at least. At the present time the volume ranges are "compressed" somewhat by adjustment of amplification under control of an operator at the pickup point. This tends to make easier the radio transmission problem as well as the wire transmission problem. The range of volume which is now delivered, as read by a "volume indicator" (a meter which roughly indicates the peaks), is of the order of 30 db (3.4 nepers), which means that during the fortissimo parts of programs the power which is transmitted is about 1,000 times as great as it is during the pianissimo portions.

The designer of the wire circuits must be concerned lest during those periods when the program power is strong, the program circuits produce an undue amount of disturbance in neighboring circuits which may be transmitting other programs or telephone messages. The designer is also concerned lest when the program power is weak the programs be unduly interfered with by noise or crosstalk from other circuits. He must particularly consider the noise and crosstalk which may be heard during pauses in programs. During such pauses it is very annoying to the listeners to hear a background of noises of various sorts and it is essential that the listeners be unable during such pauses to pick up intelligible speech from telephone message circuits crosstalking into the program circuit.

At the present time generally satisfactory results are being obtained in transmitting the volume range of about 30 db (3.4 nepers). Considerably more must be done both in the radio and in the wire systems, however, before there can be transmitted volume ranges comparable with those put out by symphony orchestras, high-grade artists, and the like.

Having indicated in a general way the requirements of program transmission circuits, there will next be described the wire systems which are now in use in the United States.

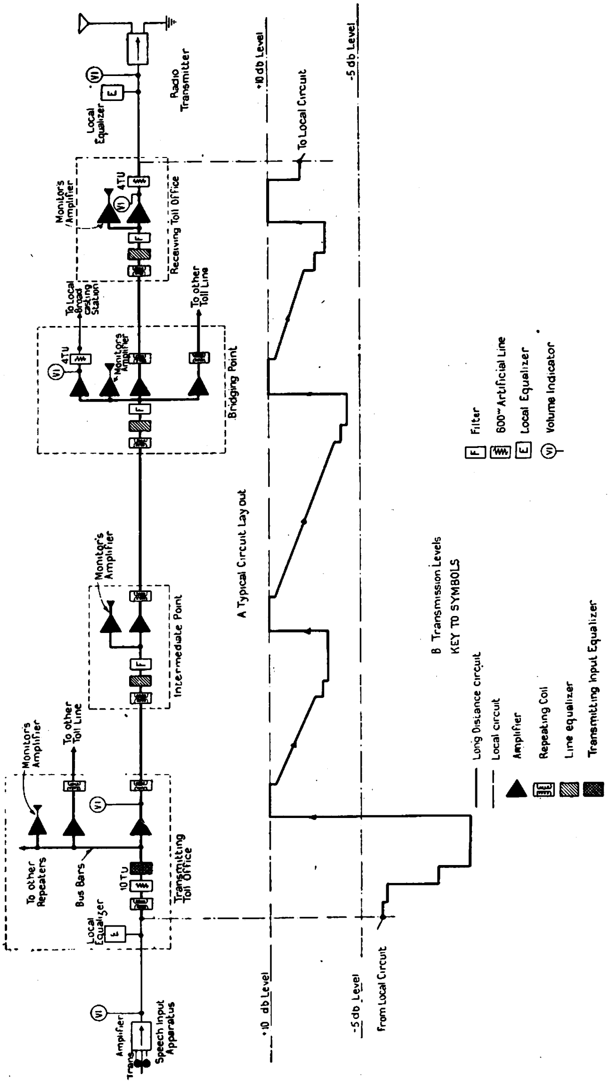

Trans.
Amplifier
Speech Input Apparatus
VI
Local Equalizer
E
10TU
Transmitting Toll Office
Bus Bars
To other Repeaters
Monitors Amplifier
To other Toll Line
Monitor's Amplifier
F
Intermediate Point
4TU
Monitors Amplifier
To Local Broadcasting Station
To other Toll Line
Bridging Point
Monitors Amplifier
Receiving Toll Office
Local Equalizer
Radio Transmitter
A Typical Circuit Lay out
+10 db Level
-5 db Level
From Local Circuit
To Local Circuit
B Transmission Levels
KEY TO SYMBOLS
Long Distance circuit
Local circuit
Amplifier
Repeating Coil
Line equalizer
Transmitting Input Equalizer
Filter
600ω Artificial Line
Local Equalizer
Volume Indicator

Fig. 3

2062

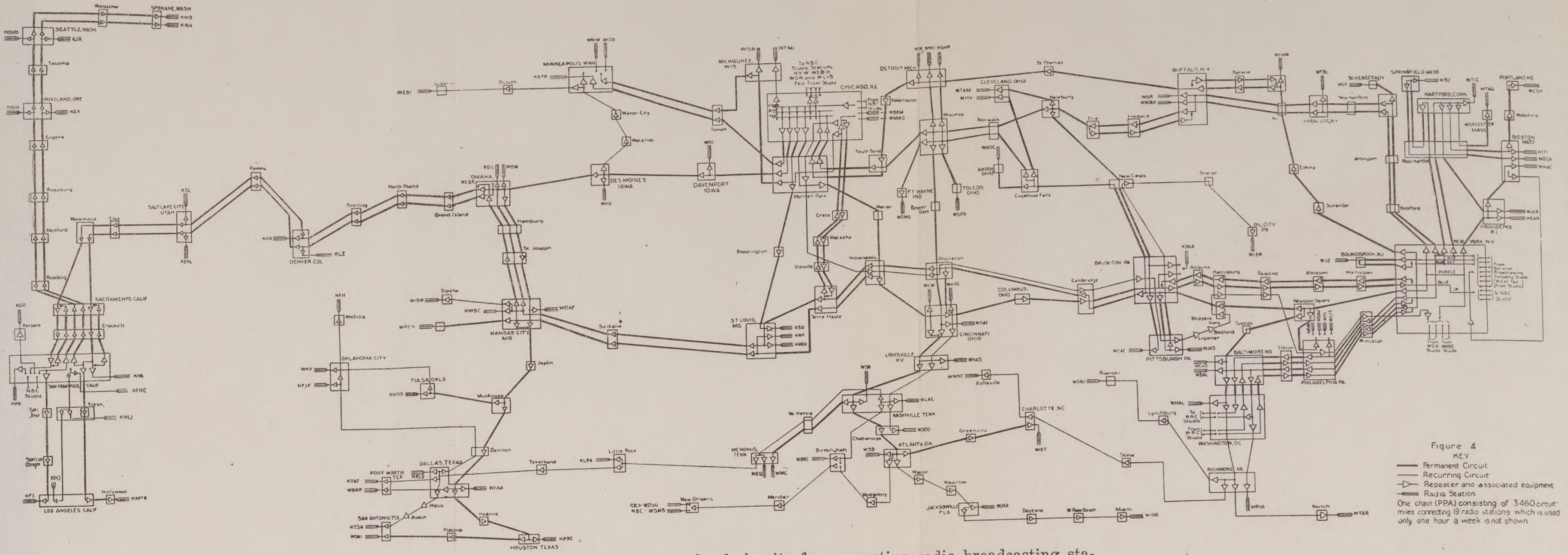

SPOKANE, WASH
KHQ
KGA
SEATTLE, WASH
KOMO
KJR
Tacoma
PORTLAND, ORE
KGW
KEX
Eugene
Roseburg
Medford
Redding
Winnemucca
Elko
SALT LAKE CITY UTAH
KSL
KDYL
Rawlins
DENVER COL
KOA
KLZ
Sterling
North Platte
Grand Island
OMAHA, NEBR.
KOIL
WOW
SACRAMENTO CALIF
KGO
Oakland
Crockett
SAN FRANCISCO, CALIF
N.B.C. Studio
KPO
KYA
KFRC
San Jose
Fresno
KMJ
San Luis Obispo
KHJ
KFI
LOS ANGELES CALIF
Hollywood
KMTR
WEBC
Duluth
MINNEAPOLIS, MINN
KSTP
WRHM
WCCO
Mason City
Waterloo
DES MOINES IOWA
WHO
DAVENPORT IOWA
WOC
Tomah
MILWAUKEE, WIS
WISN
WTMJ
Hamburg
St. Joseph
Topeka
WIBW
KMBC
WREN
KANSAS CITY MO
WDAF
KFH
Wichita
OKLAHOMA CITY
WKY
KFJF
TULSA, OKLA
KVOO
Muskogee
Joplin
Sedalia
Denison
DALLAS, TEXAS
FORT WORTH TEX
KRLD
KTAT
WBAP
WFAA
Waco
Austin
SAN ANTONIO, TEX
KTSA
WOAI
Hearne
Flatonia
HOUSTON TEXAS
KPRC
Texarkana
Little Rock
KLRA
To N.B.C. Studio Stations KYW WEBH WGN and WLIB Fed From Studio
CHICAGO, ILL
From NBC Studio
WBBM
WMAQ
Kalamazoo
South Bend
Morrell Park
Bloomington
Crete
Watseka
Danville
Terre Haute
ST LOUIS, MO
KSD
KWK
KMOX
Marion
Indianapolis
DETROIT MICH
WJR
WWJ
WGHP
Maumee
FT WAYNE IND
WOWO
Beaver Dam
TOLEDO, OHIO
WSPD
St Thomas
CLEVELAND OHIO
WTAM
WHK
Newburg
Norwalk
WADC
AKRON OHIO
Cuyahoga Falls
Phoneton
WLW
WKRC
CINCINNATI OHIO
WSAI
COLUMBUS, OHIO
Cambridge
LOUISVILLE KY
WHAS
WSM
NASHVILLE TENN
WLAC
Mc Kenzie
WDOD
Chattanooga
MEMPHIS, TENN
WREC
WMC
Birmingham
WBRC
ATLANTA, GA
WSB
Macon
Waycross
Montgomery
Meridian
New Orleans
CBS - WDSU
NBC - WSMB
JACKSONVILLE FLA
WJAX
Daytona
W Palm Beach
Miami
WIOD
WWNC
Asheville
CHARLOTTE, NC
WBT
Greenville
Selma
Roanoke
WDBJ
Lynchburg
Erie
Fredonia
BUFFALO, N.Y.
WGR
WMAK
Batavia
WHAM
SYRACUSE, N.Y.
WFBL
New Hartford
SCHENECTADY
WGY
Elmira
Scranton
Amsterdam
SPRINGFIELD, MASS
WBZ
HARTFORD, CONN
WTIC
WTAG
WORCESTER MASS
PORTLAND, ME
WCSH
Wakefield
BOSTON MASS
WEEI
WBZA
WNAC
West Hartford
Bedford
PROVIDENCE R.I.
WJAR
WEAN
New Castle
Sharon
OIL CITY PA
WLBW
BRUSHTON PA
KDKA
Altoona
Harrisburg
Reading
Allentown
Morristown
Shippensburg
Bedford
Ligonier
Tuxedo
PITTSBURGH PA
WCAE
WJAS
Newtown Square
WFAN
WCAU
WFI
WLIT
BOUND BROOK, N.J.
WJZ
NEW YORK N.Y.
PURPLE
BLUE
From National Broadcasting Company Studio (WEAF Fed From Studio)
To N.B.C. Studio
From WOR Studio
From WABC Studio
Princeton
Elkton
BALTIMORE, MD
WCAO
WBAL
PHILADELPHIA PA.
WMAL
To W.R.C. Studio
From W.R.C. Studio
WASHINGTON, DC
RICHMOND, VA
WRVA
Norfolk
WTAR
Figure 4
KEY
Permanent Circuit
Recurring Circuit
Repeater and associated equipment
Radio Station
One chain (P.P.A.) consisting of 3460 circuit miles connecting 19 radio stations which is used only one hour a week is not shown

Fig. 4—Bell System network of circuits for connecting radio broadcasting stations. Layout of permanent and recurring program transmission circuits with repeater equipment on the basis of normal direction of transmission. January 15, 1929.

The present-day program transmission circuits in the United States are "on a voice-frequency basis," which means that the waves transmitted over the circuits are essentially copies of the sound waves impinging on the microphones. Most of the circuits now being provided are carried by the familiar open wires, usually copper wires 0.165 in. (4 mm) in diameter spaced about 1 ft. (30 cm) apart on the crossarms. The transmission properties of an open-wire pair without loading are well suited for program transmission purposes since the distortion is comparatively small although it is far from negligible. Spaced at intervals on these circuits, averaging roughly 150 miles (240 km) apart, are one-way repeaters or amplifying devices. Along with these amplifiers are other electrical devices for counteracting the distortion introduced by the open-wire circuits, incidental cables involved, etc. Other one-way repeaters are provided at the terminals of the circuit. Considerable technical refinement is, of course, involved in the design of these amplifiers and of the auxiliary apparatus associated therewith which cannot be gone into here.

In setting up the program transmission circuits, an important part of the work consists in making measurements at different single frequencies within the band which it is desired to transmit over the circuit. Before making such overall measurements, the amplifiers and auxiliary apparatus are so adjusted locally as to compensate for the amount of distortion which theory and experience indicate should be expected. Then, final adjustments are made by certain specially provided adjustable parts in accordance with the overall measurements. Such overall tests and adjustments are, in general, made daily.

In setting up these circuits, another important consideration is that each amplifier carry its proper load or, in telephone parlance, each amplifier deliver to its associated line the proper output level. To insure this, diagrams are prepared in advance, showing the desired transmission levels at each repeater, a typical diagram being shown in Fig. 3. In setting up the circuits, the repeater gains are first set to values which theory and experience indicate should result in conditions as shown in the prescribed transmission level diagram. Testing current is then applied to the sending end of the circuit and sensitive measuring devices are applied at the output of each repeater. If the results of these measurements do not accord with the transmission level diagram, suitable adjustments are then made.

In building up the large chains which tie together a considerable number of radio transmitters, wire distributing centers are provided at strategic points. Fig. 4 shows the circuit layout of the various chains which have been referred to and indicates in a general way how the various chains are interconnected and arranged for switching at certain distributing centers.

In the United States the largest distributing center is, naturally, in New York City, since the bulk of the program material originates at that point. At such a distributing center a special collection of various forms of equipment is provided consisting of one-way amplifiers, loud speakers, multifrequency oscillators, various forms of transmission measuring devices and miscellaneous apparatus. The photograph of Fig. 5 shows a portion of the program layout in the New York long distance telephone office as of January 15, 1929. The various bays at the left carry the line apparatus associated with branches of various chains. In the rear are located the transmission measuring apparatus and multifrequency oscillators. In the foreground are the terminals of various telegraph order wires.

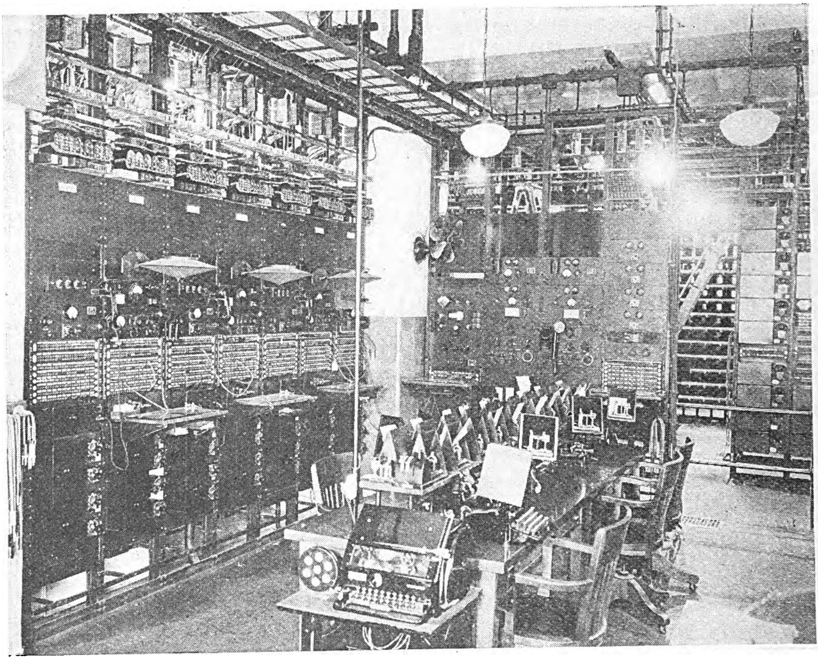


Fig. 5

In transmitting programs over a wire network, as has been pointed out above, it is important that the volume range be held within proper limits. It is one of the obligations of the one who "picks up" the program to hold his range of volume between proper limits. At the central distributing point those in charge of the wire circuits usually find it desirable to make checks from time to time to insure that the proper range of volume is maintained. This checkup is made by means of a device known as a "volume indicator" similar to the one which the program supplier uses for purposes of regulating his volume range. Other volume indicators are provided at various strategic points in the wire network in order to insure that the proper range of volume is reaching these points. In addition to regularly making these observations by means of volume indicators, loud speaker monitoring observations are continually made at practically all repeater points.

The results of these observations are transmitted back to the control points periodically by means of telegraph order wires so that the control operator knows at all times the condition of transmission at every point in his territory.

With the network chains grown to such vast proportions as indicated in Figs. 1 and 4, it is essential that the system for controlling the networks be such that all points involved be in instant communication with certain designated control points. To accomplish this, the United States has been divided into four areas, each area of which is under the control of a distributing center or control station. The four control stations in the United States at present (January 15, 1929) are, New York covering the eastern section, Chicago the western section, Cincinnati the southern section, and San Francisco the Pacific Coast section. Each of these control points is connected to every repeater point in its area by means of telegraph order wires and in addition is connected to every radio station in the area served by the networks under its control. The various control points are also connected together by means of order wires and arrangements are provided so that New York can be placed in communication with any of the radio stations in the United States which are served by the chains. The total telegraph wire mileage employed for this service is now approximately 43,000 miles (70,000 km).

A large corps of specially trained telephone men is needed to supervise properly the transmission performance of the chains as well as to take care of the switching and general coordination work involved. At present, about 300 men are employed in the United States for this service, these men, of course, being in addition to those who care for the regular wire and equipment maintenance.

ACKNOWLEDGMENT

Acknowledgment is made to H. S. Hamilton for considerable assistance in connection with the preparation of the text and particularly of the drawings, and to G. S. Bibbins and H. C. Read for furnishing most of the statistical data.

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RADIO BROADCASTING REGULATION AND LEGISLATION*

By
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GOVERNMENTAL regulation of radio has been, and still is, a difficult problem. This is true throughout the whole realm of radio, conspicuously so in broadcasting. In many countries the solution is being sought through a government monopoly of broadcast transmission, in the United States through private ownership and operation. In every country there is strict government control, the tendency in most cases being to vest this control in a specially created commission.

The legislative problem arises from certain inherent limitations of radio. These are primarily two: (a) radio waves spread out everywhere, stopping at no boundaries, and capable of great mutual interference; (b) the number of radio communication channels is definitely and severely limited. The implications of these basic principles are far-reaching. The first has led to international regulation, now embodied in the 1927 International Radio Convention.

The second (the limited number of communication channels) is the essential problem with which the governments of the various nations are wrestling. The right to operate radio transmitting stations has become so valuable that the demand for channels in most countries far exceeds the supply. The number of channels is being increased from time to time, as radio technique develops, but at any given time there are only a limited number of communication channels. If stations are permitted to operate in excess of the capacity of these channels extensive interference develops and radio operation becomes unsatisfactory. In the present state of radio technique the number of channels in the whole radio spectrum is as follows: in the low frequencies (15 to 550 kc), there are something less than 1000 channels; in the principal broadcast band of frequencies (550 to 1500 kc), there are 96 channels; and in the high frequencies (from 1500 to 23,000 kc), there are about 1260 channels. Thus we have a total of about 2350 channels available for the radio traffic of the entire world. On some of these frequencies, particularly the very low and the very high, it is not feasible to have more than one station operating at any given time in the whole world. On many of the others a large number of stations can operate simultaneously. The facts regarding these frequencies are well enough established to permit us to calculate the number of stations which can be operated without causing excessive interference.

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The fact that the number of channels is limited and the number of stations assignable to any one channel is again limited, imposes upon governments the necessity of choice among applicants for the radio channels. This is the underlying reason why radio commissions have come to be created. To provide for choice among those who aspire to construct and use radio stations, the United States Government created not only a commission as its instrument to make the choice, but also a judicial principle to be the basis of choice, viz., the principle of "public interest, convenience, or necessity."

The interpretation and application of this phrase is a novel development in jurisprudence. The phrase is borrowed from legal terminology used in the regulation of public utilities, such as street-car lines and gas companies. Radio broadcasting stations are not public utilities, and yet the test of a public utility must be applied to them for the two basic reasons already given, viz., the potential interference of each radio transmission with every other, and the fact that the number of channels is sharply limited.

The test of "public interest, convenience, or necessity" is being applied in the United States to mean that the rights of the listeners are superior to those of the broadcasting stations. This means that, as far as possible, interference must be avoided. It means that rural listeners remote from any station, as well as city listeners, must be given service. It means that excessive duplication of programs by many stations cannot be permitted, and that high-power stations cannot be located in the midst of cities. Perhaps the most important implication is that the total number of broadcasting stations must be limited to the number necessary to prevent undue interference.

The number of stations which can be in simultaneous operation depends on a number of engineering factors. Chief among these is the width of channel necessary for each. Extensive experience has shown that the minimum separation between adjacent channels, for satisfactory musical reception, is 10 kc. Even this is not quite sufficient for musical reproduction of the highest quality. In the frequency band from 850 to 1500 kc (both inclusive) there are a total of 96 available frequencies. If there is more than one station, using in excess of 1-kw power, on one frequency in an area the size of the United States, interference is likely to result. Now there are 620 stations in the United States, and the difficulty of regulating their operation is at once apparent.

In order to minimize interference among this large number of broadcasting stations it has been necessary to: (a) limit the simultaneous operation of stations by making many of them divide time; (b) assign frequencies carefully selected with regard to geographical separation of stations, to reduce interchannel interference (i.e., disturbance of reception of a station on one frequency by other stations on adjacent frequencies); and (c) limit the power of stations so they would not cause interference to other stations on the same frequency.[1] The most striking of the problems involved in the new allocation was the carrying out of requirement (c) just mentioned. Stations assigned to the same frequency have not, up to the present, been able to maintain their frequencies with sufficient accuracy to prevent the existence of a slight difference or beat frequency, producing what is commonly known as heterodyne interference, or whistles. Unfortunately the heterodyne interference reaches out to enormously greater distances from a station than the program. Consequently the operation of two or more stations on a channel results in an area of destructive interference far in excess of the area in which program service is provided. For instance, a 5-kw station's program can be heard with fair intensity under good conditions at 100 miles while the heterodyne interference from two such stations is heard at 3000 miles. Two stations of 5 kw or more, therefore, cannot be assigned the same frequency in the United States. It is possible, on the average, to put two or more 1-kw stations on the same frequency if they are at least 1800 miles apart, and two or more 1/2-kw stations if they are at least 1200 miles apart. All stations subject to these restrictions have only a small service area, and give no service to remote rural areas. Such distant service is given only by stations having exclusive use of the channels to which they are assigned.

In order to provide rural service 40 channels are each used by one station exclusively. The stations on the exclusive channels not only serve very great areas but deliver a more satisfactory intensity at every point within those areas. Their service is better for all concerned, the greater the power they use. This fact is not commonly understood by others than radio engineers. It is clear when the distinction between the exclusive and the other channels is comprehended. Service on the non-exclusive channels would be utterly ruined if the power limits fixed by the facts of heterodyne interference should be exceeded, and in consequence such stations cannot in general use more than 1 kw. But on the exclusive channels the service is better the higher the power level, and indeed such stations will not be serving the public most effectively until the level reaches hundreds of kilowatts.

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[1] For a detailed description of this allocation from the engineering viewpoint, see "Analysis of Broadcasting Station Allocation," PROC. I. R. E., **16**, 1477–1485; November, 1928.

There is some hope that the limitation of power and service of the non-exclusive channels may be overcome. If the frequencies of stations on the same channel are maintained to a certain very high accuracy, the heterodyne or whistle becomes inaudible. The technique of frequency control is fast approaching this goal and success has been attained in isolated instances. The satisfactory service area of such "synchronized" stations is not yet known, but it is believed that it will be substantially greater than when heterodyne interference exists. The significance of this is that the present power limits for stations on shared channels can be raised, better service given, and wider areas served. Synchronization is therefore looked for as the next great advance in broadcasting.

This discussion of broadcasting has been largely with reference to night conditions. Broadcast transmission is entirely different in the daytime. Transmission distances are much less, and somewhat greater power can be allowed the stations. Furthermore, additional stations are licensed for daytime operation only.

The difference between day and night transmission conditions raised one technical problem of considerable moment, viz., determination of the time when day ends and night begins, and thus at what hour daytime stations should close. Investigation revealed that the change from day to night radio conditions extends over a period of something more than an hour and a half, beginning about a half hour before sunset and closing an hour after sunset. The most reasonable time to choose as the transition point is the moment of sunset.

There are many other technical problems involved in the regulation of broadcasting. The limitations of this brief discussion will permit a mere listing of examples, such as chain program limitation; visual broadcasting (television); the requirement of an artificial (dummy) antenna in broadcasting stations for use during warming-up periods; the location of high power stations with respect to populous areas; requirement of highly accurate frequency control; allowable ratio of day to night power; permissible intensity of harmonics; percentage modulation; and fidelity of transmission.

Radio legislation and regulation have the peculiar difficulty that the facts dealt with are extremely complex. They are indeed rapidly shifting. Not only must allowance be constantly made for the flux of changes inherent in a rapidly developing art, but radio waves themselves exhibit extraordinary vagaries. Orderly radio regulation must proceed on a consideration of the distances at which the waves are received. But distances vary enormously between day and night, from season to season, even from night to night, and are different over different kinds of terrain. Knowing this is not to counsel despair. These vagaries have, after all, certain discernible laws becoming more and more calculable as the results of scientific investigations accumulate. It is not necessary to throw up our hands and say that the whole situation is chaotic. In spite of their vagaries, radio phenomena are subject to known engineering principles. Violation of such engineering principles in radio would sooner or later reduce the service of radio to the public.

Summarizing, the regulation of radio broadcasting involves extensive and difficult problems. These arise largely from certain outstanding facts or principles. First, radio waves spread out everywhere and potentially interfere with one another. Second, the available number of communication channels is definitely limited. Another controlling principle, as the art stands today, is that heterodyne interference sharply limits the power that may be permitted any two or more broadcasting stations on the same channel. Finally, radio wave transmission is characterized by extreme vagaries. The facts and implications of each of these principles are subject to constant revision as radio progresses. Such facts constitute the natural limitations of radio regulation and legislation.

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MEASUREMENT OF FREQUENCY*

BY

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Summary—*Absolute measurement of frequency.—Various methods of absolute measurement of frequency are first described. Particular attention is devoted to the stroboscopic method and the phonic motor method for audio frequencies. A modified method of harmonic comparison is also presented.*

Standard frequency oscillators.—The author has designed a clock-controlled oscillator which consists of a tube-driven 25-cycle adjustable tuning fork and a chronometer. The performances of various kinds of tuning fork oscillators are discussed mathematically, and a new type, which has a magnetic device of such nature as to make the electromagnetic controlling force extremely small, is described. The factors affecting the constancy of frequency of quartz oscillators are discussed.

Resonators.—Several types of electrical and mechanical resonator are compared with respect to their usefulness as frequency standards. For expressing the sharpness of resonance, the author recommends the adoption of the quantity Λ *defined by the equation* $\Lambda = \omega_0/2\delta$, *where* ω_0 *is the angular velocity at resonance and* δ *the damping constant. Experimental results are presented showing the sharpness of resonance of a number of resonators of different types.*

I. INTRODUCTION

AT the present time the precise measurement of frequency is becoming more and more important. In this paper, first the absolute measurement of frequency, second the primary standards of frequency, and finally the behavior of working standards are described.

Since frequency is the reciprocal of time, the standard of time and the measurement of time intervals will be the most important factors throughout the present problem. As the primary standard of time, the author has used a Riefler standard clock which has been corrected against the standard clock at the astronomical observatory by means of radio signals. As the working standard of time, a Nardin chronometer has been used. The time intervals are measured by the chronometer and a tape chronograph as shown in Fig. 1. By this device, a measurement extending over six minutes will enable us satisfactorily to obtain an accuracy of time interval of the order of one part in a hundred thousand.

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II. ABSOLUTE MEASUREMENT OF FREQUENCY

1. Stroboscopic Method

Radio frequencies are usually measured by means of harmonic comparison with an audio-frequency standard. Hence first of all the absolute measurement of audio frequency must be considered.

The author recommends the stroboscopic and the phonic motor methods as the most precise means of measuring audio frequency. As already known, the measuring device in the stroboscopic method consists of a sector revolving at a constant speed and a neon glow lamp illuminating it, as shown in Fig. 2. If the frequency of the voltage impressed on the glow lamp is equal to that of the revolution of the pattern in the sector, the pattern looks as if it stood still. When the pattern moves so slowly that the rate of its movement can be counted, the frequency f of the voltage can be written as

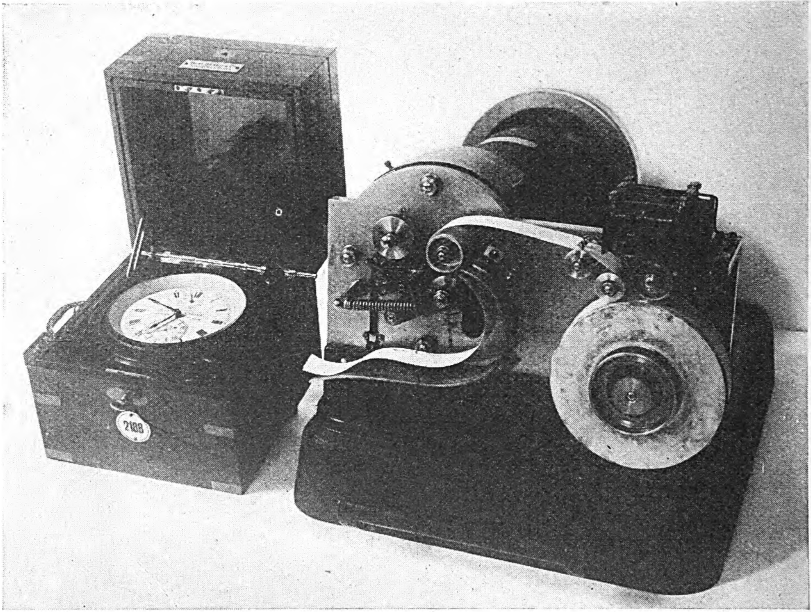

Fig. 1

$$f=f_m \cdot N \pm n/T \tag{1}$$

where f_m is the number of revolutions of the disk per second, N the number of patterns on the disk, T the measured time interval required for n transits of the pattern. n/T is negative when the direction of displacement of the pattern coincides with that of the sector.

The most important point in the stroboscopic method is the constancy of speed of the sector disk. There are many devices for constant speed governors. The author has succeeded in developing a highly sensitive speed governor, shown in Fig. 3.

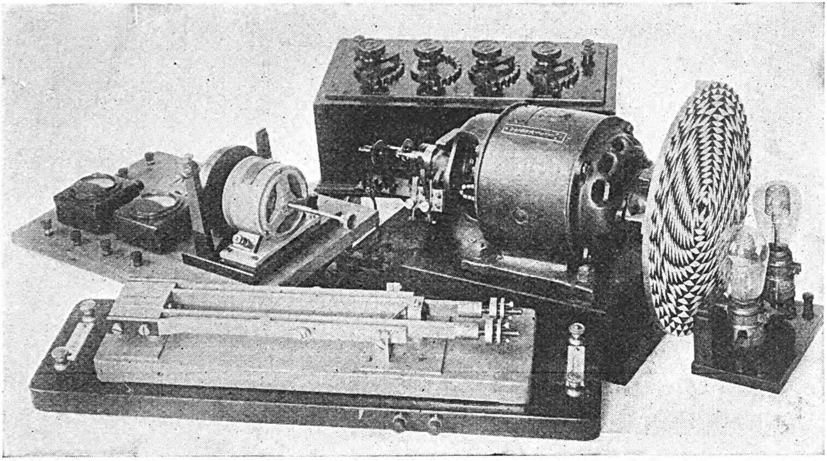

Fig. 2

The author's method is somewhat similar to that of R. H. Hough and F. Wenner[1] in principle, but it presents a remarkable improvement. The principle of this governor consists in the synchronism of a motor with the vibrations of a tuning fork. In the new device, a half-ring commutator is on the shaft of a d-c motor which is separately excited. Two brushes on this commutator are connected to the terminals of a non-inductive adjustable resistance r in the field circuit of the motor through the commutator brushes of a phonic motor, which consist of wires stretched tangentially on the commutator of the phonic motor.

The latter commutator consists of ten segments and its diameter is 10 cm, so that it acts as a flywheel to prevent the hunting of the phonic motor. The phonic motor and the tuning fork of 25 cycles per sec. are of the Tinsley-Wood type. Now if owing to any cause the speed of the motor is slightly decreased, the time interval for short-circuiting the resistance r decreases, and consequently the field current decreases until it is of the proper value to keep the motor from falling below synchronism.

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[1] R. H. Hough and F. Wenner, *Phys. Rev.*, **24**, 535, 1907.

Since in this device commutators are used for producing the electrical contacts, the time interval for short-circuiting the resistance is very reliable, and consequently the performance of the governor is sufficiently precise. It is ascertained experimentally that the governor is most sensitive when the time interval for short-circuiting the resistance is 1/4 of the period corresponding to one complete revolution of the motor, as shown in Fig. 4. The resistance *r* is 285 ohms, and the exciting voltage 90 volts.

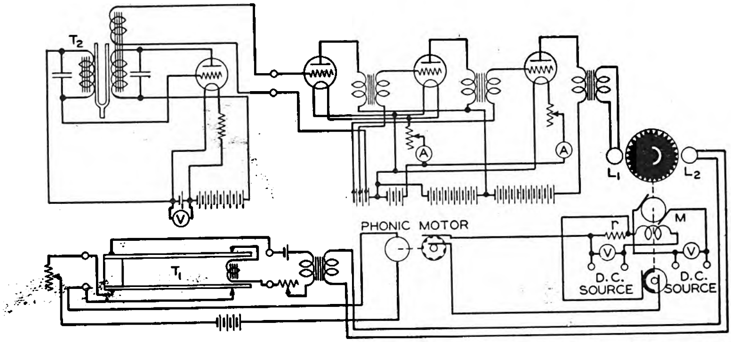

Fig. 3—Apparatus and circuits for stroboscopic comparison of frequencies.*

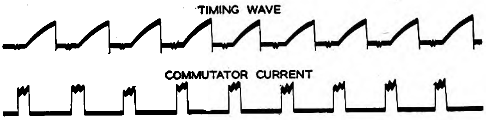

Fig. 4

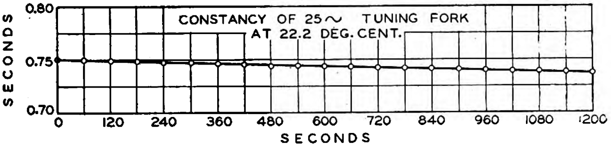

Fig. 5

* In Fig. 3, T_1 is the special 25-cycle make-and-break fork mentioned in the text, and T_2 the fork with which it is to be compared. T_1 consists of two flat metal springs fastened to an insulating block. The upper contact is the make-and-break in the circuit that drives the fork, while from the lower contact impulses are sent for driving the phonic motor. The make-and-break wheel on the latter is in series with the half-ring commutator on the shaft of the d-c motor M, and together they regulate the field excitation of M in such a way as to hold the speed constant. The stroboscope is on the shaft of M. The neon lamp L_1 is supplied with alternating current amplified from the output of the tuning fork oscillator T_2. The lamp L_2 is lighted at the frequency of T_1. Thus by means of L_2 and the stroboscope disk a check is obtained of the synchronism between T_1 and M while L_1 serves for comparing T_2 with the speed of the disk and hence with T_1.—*Ed.*

The constancy of the tuning fork is an important consideration in this device. It is investigated by comparing two dots on the tape of the chronograph shown in Fig. 1, one of which is made at the rate of one per second by an electric contact on the phonic motor operated by the tuning fork, and the other by the chronometer. The experimental result obtained during a run of 20 minutes, shown in Fig. 5, shows that the constancy of this fork is sufficient. It will be noticed that the waveform of the voltage on the glow lamp must be steep to obtain a clear pattern.

By this stroboscopic method with the above speed-governor, the author has succeeded in obtaining a precision of one part in a hundred thousand in the absolute measurement of frequency.

In this method, it is an excellent feature that any multiple of 25 cycles, from 50 to 1025 cycles per sec., can be easily measured. The author believes that this governor will prove useful for various applications.

2. Phonic Motor Method

The phonic motor was designed by Latour and Lord Rayleigh. A high-frequency motor operated by a small power tube has been described by D. W. Dye and recently by J. W. Horton and W. A. Marrison.[2] The author has designed a phonic motor which has 50 teeth and two pairs of poles. This motor can be operated by a UX171 tube from 100 to 700 cycles per second. For frequencies over 700 cycles, a synchronous motor made by the General Radio Co. is now being used, until a new high-frequency motor is finished. In the phonic motor method, the hunting of the motor is noticeable. To prevent this hunting, mercury is frequently used, but by the stroboscopic method it is found that mercury is not sufficient for this purpose. In this connection, it may be added that it is most important that the pole arrangement be as regular as possible. However, since perfection is not to be expected, the motor must be run long enough to eliminate the effects of hunting from the experimental result.

By this method, we may expect an accuracy as high as one part in a hundred thousand in frequency.

3. Harmonic Comparison Method

As is well-known, a radio frequency can be measured by harmonic comparison with a standard frequency. There are many methods of accomplishing this, such as the multivibrator and the cathode ray oscillograph.

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[2] J. W. Horton and W. A. Marrison, PROC. I. R. E., 16, 137; February, 1928.

The author has developed a method of harmonic comparison, in which the measuring device consists of a multifrequency oscillator, a selector circuit and a frequency bridge. The multifrequency oscillator is somewhat similar to the timing device in Gabor's cathode ray oscillograph.[3] By adjusting the resistance in the plate circuit so as to make the time constant of the plate circuit nearly equal to the reciprocal of the circular frequency applied to the grid of the valve, the desired result can be expected. The oscillating current contains an ample supply of the higher harmonics.

The selector circuit consists of a fixed air condenser, a variable air condenser and a low loss coil, and its sharpness of resonance is sufficiently large.

The resonant sharpness of the frequency bridge is so great as to give a balance within two tenths of a cycle per second from 100 to 1,500 cycles, and the bridge is calibrated by the phonic motor method. In this method, the frequency to be measured can be easily compared with the 50th harmonic of the audio-frequency standard.

High frequencies over 50 kc can be similarly compared with harmonic frequencies of a quartz oscillator, the fundamental frequency of which can be determined by the above method, while in this case the multifrequency oscillator is not needed. In the method mentioned above, we can expect the accuracy of one part in a hundred thousand for all frequencies greater than 20 kc.

III. STANDARD FREQUENCY OSCILLATORS

1. Introduction

The absolute measurement of frequency is somewhat laborious, so that in order to preserve the results of absolute measurement a standard frequency oscillator is needed. The standard frequency oscillator must fulfil the following conditions:

1. Frequency must not depend on external circumstances,
2. Permanence of frequency.

The oscillators satisfying the above conditions are the following:

1. Clock-controlled oscillator,
2. Tuning-fork oscillator,
3. Quartz oscillator.

First, we will consider the performance of a homogeneous mechanical vibrator.

There are to be considered the longitudinal, lateral (or flexural), and torsional modes of elastic vibration. If we consider a thin bar with a rectangular cross section, and if the thickness, length and breadth are x, y and z, respectively, then the natural frequency of longitudinal vibration can be expressed by

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[3] D. Gabor, "Forschungshefte d. Studiengesellschaft für Hochspannungsanlagen, September, 1927.

$$f_{lx} = mv_l/2x, \quad f_{ly} = mv_l/2y, \quad f_{lz} = mv_l/2z \tag{2}$$

where m is any integer and v_l denotes the propagation velocity $\sqrt{E/\rho}$, while E is the modulus of elasticity, and ρ the density.

For the flexural vibration of a Curie cut plate in the yz plane, we have

$$f_b = \frac{k^2 v_l}{4\pi\sqrt{3}} \cdot \frac{z}{y^2} \tag{3}$$

where $k = (m + 1/2)\pi$, $v_l = \sqrt{E/\rho}$, and m is a constant.

For the torsional vibration, about the y-axis,[4] provided that r is less than 1/3, where $r = x/z$, we have approximately

$$f_t = \frac{mv_t}{2y} \cdot \frac{2r}{\sqrt{1+r^2}} \cdot \sqrt{1 - 0.63r} \tag{4}$$

where v_t is $\sqrt{T/\rho}$ and T the torsional rigidity. The propagation velocity depends on the physical properties of the material. The approximate values for steel and quartz are shown in Table I.

TABLE I

Material	v_l (cm/sec)	v_t (cm/sec)
Steel	$4,2\text{–}5,2.10^5$	$3,2.10^5$
Quartz	$5,4.10^5$	$3,5.10^5$

The tuning fork is a clamped-free bar, and according to Lord Rayleigh we have

$$f_b = 845900 \cdot z/y^2. \tag{5}$$

The vibration of a monochord is not an elastic wave, still its natural frequency is similarly obtained:

$$f_c = v_c/2y \tag{6}$$

and

$$v_c = \sqrt{W/\rho}$$

where W is the tension, y the length and ρ the mass of the monochord per cm.

The above equations may not be sufficiently accurate for practical cases, because actually the vibrations are somewhat complicated and very difficult to solve mathematically.

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[4] Giebe and Scheibe, *Zeitschr. f. Phys.*, **46**, p. 638, 1928.

2. Clock-Controlled Oscillator

A. Dey and C. V. Raman[5] have already shown that the vibrations of a mechanical vibrator, whose frequency is f/m, where m is any integer, can be maintained by an alternating current of which the frequency is f cycles. The present oscillator is somewhat similar to the above oscillator in principle.

The present oscillator consists of a chronometer and tube-maintained tuning fork as shown in Figs. 6 and 7.

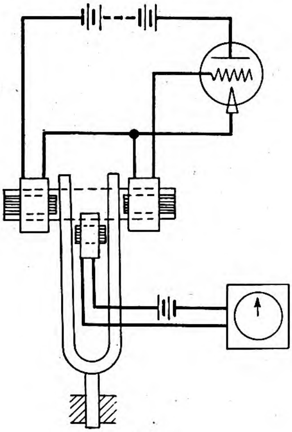

Fig. 6

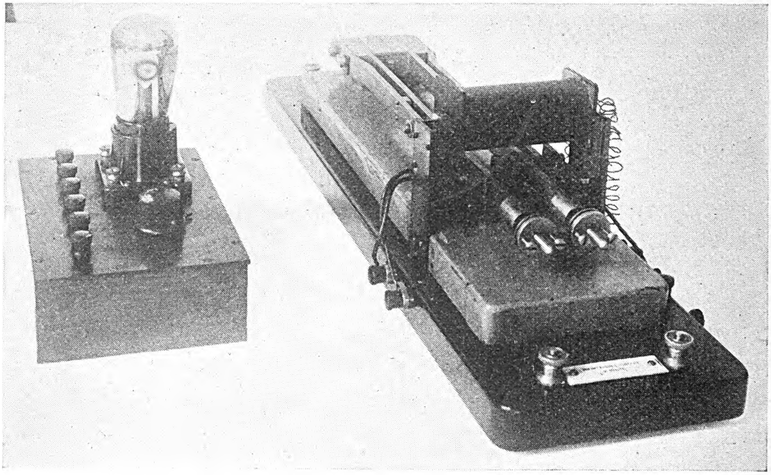

Fig. 7

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[5] A. Dey and C. V. Raman, *Proc.* Roy. Soc., 95, 533, 1919.

In the tube-maintained tuning fork, the working resonant sharpness can be extremely large, so that forced oscillations of the fork may be produced by the harmonic current, which is usually very small, over a rather wide range of natural frequency of the fork.

In Fig. 8 an adjustable tuning fork having a normal frequency of 25 cycles per second is used; f_t denotes the frequency of oscillation for various adjustments when controlled by the chronometer, and f_f the frequency without the chronometer. According to this figure, we see that the automatic synchronization can be held over a considerable range.

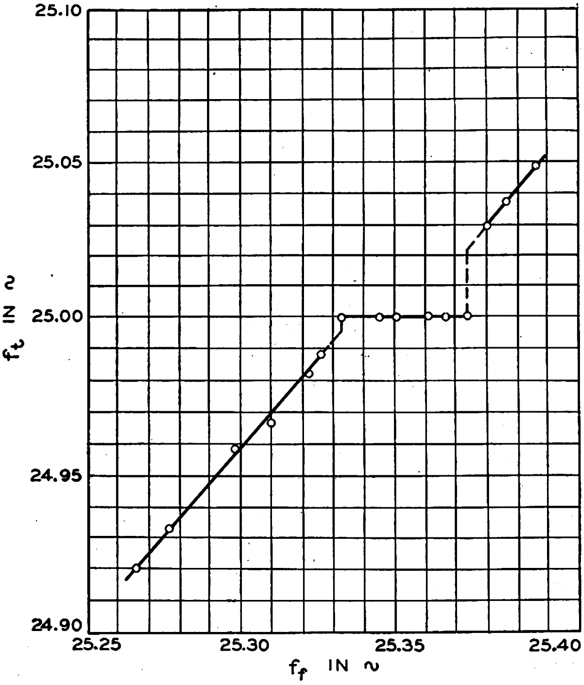

Fig. 8

In this device, it is true that the synchronization must always be produced artificially, but there is the advantage that synchronization can be easily obtained, and it is not disturbed by the surrounding temperature and other factors.

3. Tuning Fork Oscillator

The principle of the tuning fork oscillator was discovered by W. H. Eccles. The performance has been studied by S. Butterworth and T. G. Hodgkinson theoretically, and by D. W. Dye experimentally.[6]

The author considers the important factors affecting the oscillating frequency to be (a) the surrounding temperature, (b) magnetization of the fork, and (c) the clamping device of the fork. These factors will be treated as follows.

(a) *Temperature coefficient.* The temperature coefficient of an ordinary steel fork is about 1×10^{-4}; but if it be made of elinvar (Ni, 35 per cent; Cr, 12 per cent; Fe, 53 per cent), this coefficient becomes as small as -1×10^{-5}.

The temperature coefficient of an elinvar fork, made by the Sullivan Co., is measured by the stroboscopic method and by the phonic motor method. The fork oscillator is kept at a constant temperature in a bath with a sensitive thermostat for six hours, and then the frequency is measured. The experimental results are shown in Table II.

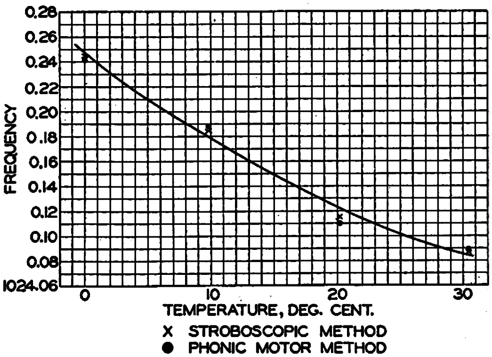

Fig. 9

TABLE II

Correction of Riefler clock	Temperature	Humidity	Measured Frequency	
			by stroboscopic method	by phonic motor method
+0.49	30.4 deg. C	66 per cent	1023.08$_7$	1024.08$_9$
+0.53	20.2	72	1024.11$_5$	1024.11$_0$
+0.49	9.9	51	1024.18$_6$	1024.18$_9$
+0.52	0.0	30	1024.24$_5$	1024.24$_0$

The two values are in sufficiently close agreement with each other, and the accuracy of measurement can be about one part in a hundred thousand. The performance curve of the first of these forks is shown in Fig. 9. A similar curve was obtained with the other fork. From these results, we have

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[6] S. Butterworth, *Proc.* Phys. Soc. (London), **32**, 345, 1920; T. G. Hodgkinson, *Proc.* Phys. Soc. (London), **38**, 24, 1925; D. W. Dye, *Proc.* Roy. Soc., **103**, 209, 1923.

$$f=1024.24_8\{1-7.6\times10^{-6}T+7.8\times10^{-8}T^2\} \qquad (7)$$

and

$$f=960.26_9\{1-12.8\times10^{-6}T+6.5\times10^{-8}T^2\}$$

Therefore, we know that the temperature coefficient of the elinvar fork has always a negative sign, and that it is nearly 1×10^{-5}.

(b) *Magnetization of fork.* We first consider the theory of the fork oscillator, which is of the untuned type.

The motion of the fork can be expressed by the following equation:

$$\frac{d}{dt}\left(\frac{\partial T}{\partial \dot{y}}\right)+\frac{\partial F}{\partial \dot{y}}+\frac{\partial V}{\partial y}=\frac{\partial T_e}{\partial y}, \qquad (8)$$

where T denotes the kinetic energy, F the dissipation energy, V the potential energy, T_e the electrokinetic energy, and y the displacement of the fork.

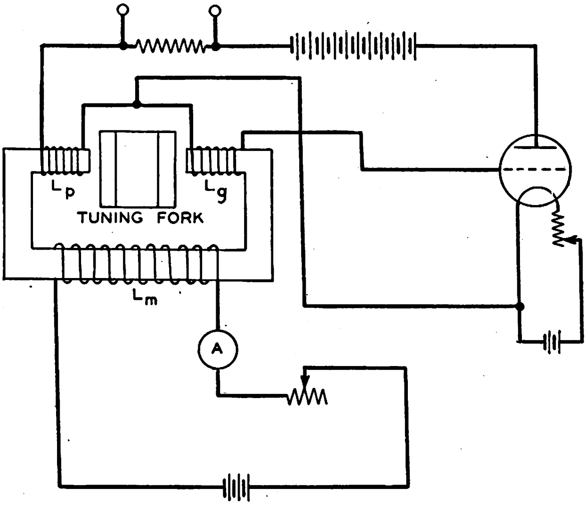

Fig. 10

Now we will suppose that the fork has one degree of freedom, and that the mechanical restoring force of the fork is not proportional[7] to the displacement y. We then have

$$T=\tfrac{1}{2}\alpha\dot{y}^2, \quad F=\tfrac{1}{2}\beta\dot{y}^2, \quad V=\tfrac{1}{2}(\gamma-\tau y^2)y^2.$$

Therefore we have

$$\{\alpha D^2+\beta D+(\gamma-2\tau y^2)\}y=\frac{\partial T_e}{\partial y}.$$

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[7] Rayleigh, "Theory of Sound," Vol. 1, p. 77.

As shown in Fig. 10, let L_p, L_g, L_m be the self inductances of plate coil, grid coil, and magnetizing coil, respectively.

Now we suppose that L_p, L_g, M_{pg} are independent of y and that

$$L_m = l_0 + l_1 y + l_2 y^2 \tag{9}$$

$$M_{pm} = M_{gm} = m_0 + m_1 y + m_2 y^2 .$$

Let I_0 be the magnetizing current, I the induced alternating current in the magnetizing coil, then

$$\frac{\partial T_e}{\partial y} = \tfrac{1}{2}(l_1 + 2l_2 y) I_0^2 + (l_1 + 2l_2 y) I I_0 + (m_1 + 2m_2 y) I_0 I_p .$$

Considering the saturation of the tube, we have

$$I_p = G_v E_g (1 - g E_g^2)$$

where I_p, E_g denote alternating components, and G_v is the mutual conductance of the tube. Then we have

$$I_p = G_v (m_1 + 2m_2 y)(1 - g m_1^2 I_0^2 (Dy)^2) I_0 Dy$$

Let G_m be the conductance of the magnetizing coil, then

$$I = -G_m (l_1 + 2l_2 y) I_0 Dy$$

Therefore, the equation of motion can be written

$$\left\{ \alpha D^2 + \beta_e D + \gamma_e \right\} y = \tfrac{1}{2} l_1 I_0^2 \tag{10}$$

where $\beta_e = \beta + \beta_m - \beta_v$.

$$\beta_m = G_m (l_1 + 2l_2 y)^2 I_0^2 , \quad \beta_v = G_v (m_1 + 2m_2 y)^2 (1 - g m_1^2 I_0^2 (Dy)^2) I_0^2 ,$$
$$\gamma_e = \gamma - l_2 I_0^2 - 2\tau y^2 .$$

The above differential equation cannot be completely solved, because β_e, γ_e are functions of y. We will consider that actually the mean values of β_e, γ_e during one complete vibration act on the fork, assuming $y = Y \cos \omega t$. We can then obtain the following approximate solution:

$$\omega \doteqdot \omega_0 \left\{ 1 - \frac{l_2 I_0^2}{2\gamma} - \frac{\tau}{2\gamma} Y^2 \right\} \tag{11}$$

$$Y \doteqdot \frac{\sqrt{2}}{\omega_0 m_1 I_0 \sqrt{g}} \left\{ 1 - \frac{\beta}{G_v m_1^2 I_0^2} - \frac{G_m}{G_v} \left(\frac{l_1}{m_1} \right)^2 \right\}^{1/2} \tag{12}$$

supposing l_2/m_1, m_2/m_1 to be very small. ω_0 denotes 2π times the natural frequency of the fork, viz. $\sqrt{\gamma/\alpha}$. From the above theory, we conclude that

(1) The oscillating frequency depends on the magnetizing current; namely the frequency increases with decrease of magnetization of fork, because the electromagnetic controlling force, in addition to the mechanical restoring force, acts on the fork.
(2) The amplitude of vibration is small when the natural frequency of the fork is large.
(3) The presence of the magnetizing coil causes large damping.

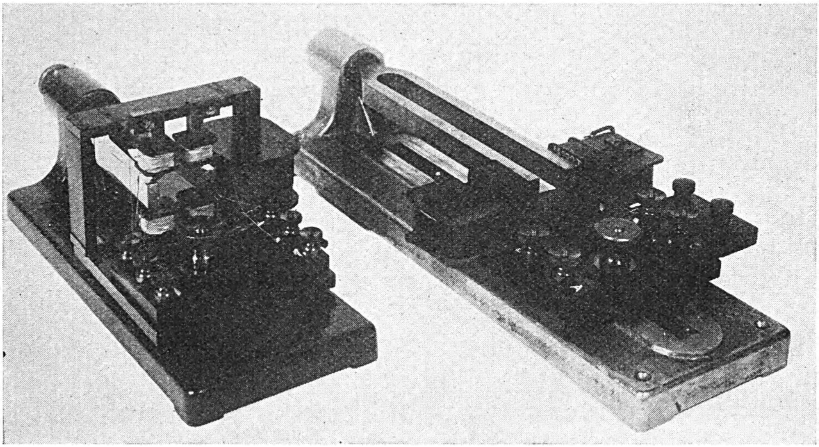

Fig. 11

The first point is most noticeable. According to the author's theory, the electromagnetic controlling force can be eliminated by making the coefficient l_2 equal to zero. In order to realize this, the author has contrived a new form of fork oscillator as shown on the left hand side of Fig. 11 while that on the right hand side is one of ordinary type.

In this new oscillator, the four poles are arranged against the fork and the poles are shaped in such a manner as to make the above coefficient l_e become zero. The relation between magnetizing current and amplitude of fork or frequency was investigated experimentally. In this experiment, a tuning fork of 1000 cycles is used and it is installed in a bath with a thermostat, whereby the temperature is always kept constant. The amplitude of the fork is measured by a cathetometer, and the frequency of oscillation by the phonic motor, chronometer and chronograph already mentioned. These experimental results are shown in Fig. 12. Fig. 12A is for the ordinary type of fork, with tube circuit untuned, while Fig. 12B is similar but with tube circuit tuned to fork frequency. Figs. 12C and 12D were obtained with the author's new device, the tube circuits being untuned in Fig. 12C, and tuned in Fig. 12D. In the ordinary device, there is a considerable variation of frequency due to the change of exciting current, while in the new device, this variation is extremely small.

The author believes that this new device will be valuable as a standard frequency oscillator, because, even though the permanent magnet in the oscillator may become partially demagnetized after long running, the frequency is not affected.

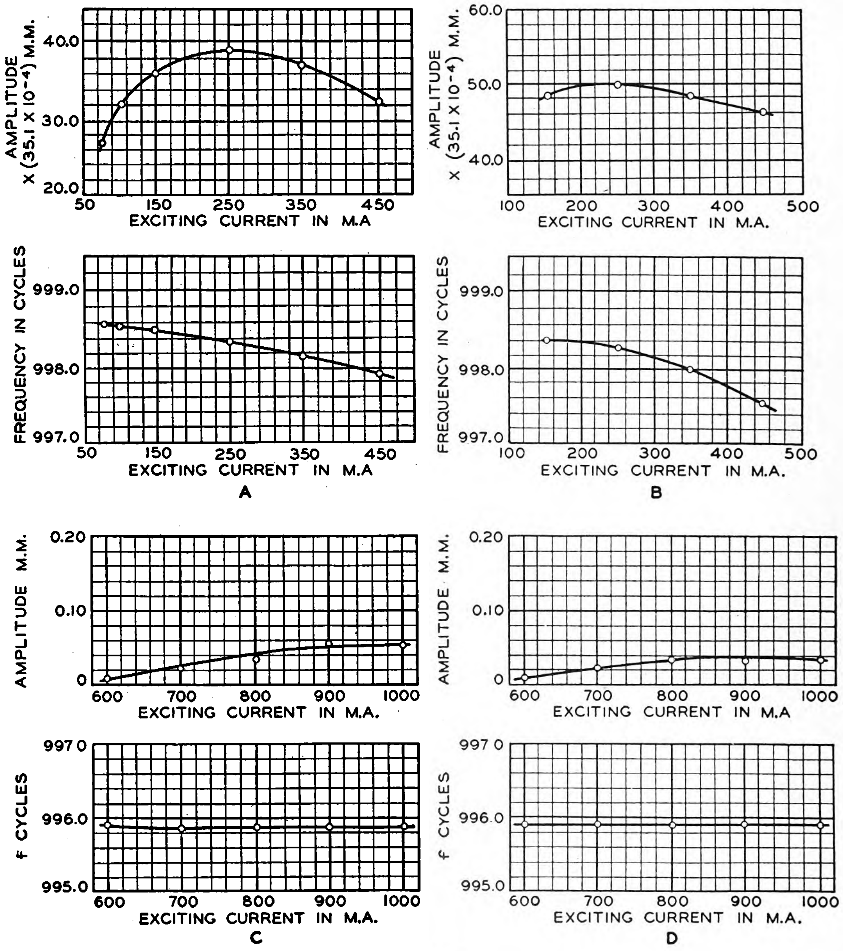

Fig. 12

Oscillators may be classified into tuned and untuned types. The author suggests that the former type is suitable for high-frequency forks, the latter for low. The monochord oscillator designed by the author may also be available as a standard frequency oscillator, but its reliability is not so great as that of the chronometer, fork and quartz.

(c) *Clamping device of fork.* Any insecurity in the clamping of the fork will introduce a considerable change in frequency. However, this may easily be removed by sufficiently clamping the fork to the base. Besides this factor, there are several others, such as weight of base, circuit conditions, etc., but these factors will not introduce any noticeable influence on the frequency. The standard frequency oscillator must be always kept in a definite state, under, as constant conditions as possible.

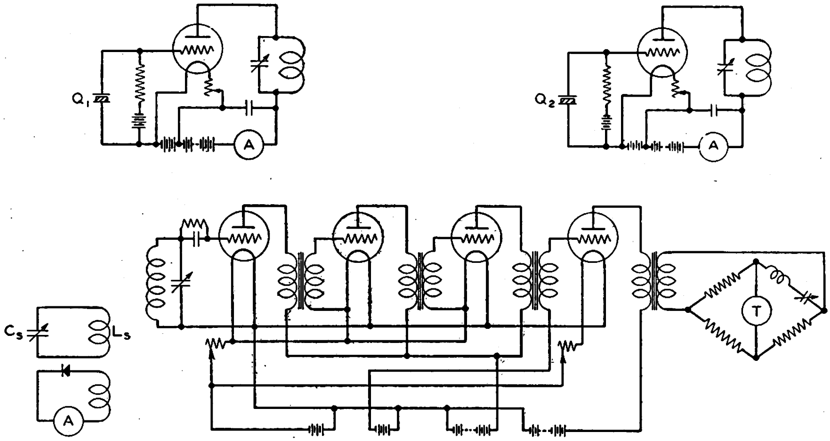

Fig. 13—Circuits for comparison of two piezo oscillators.

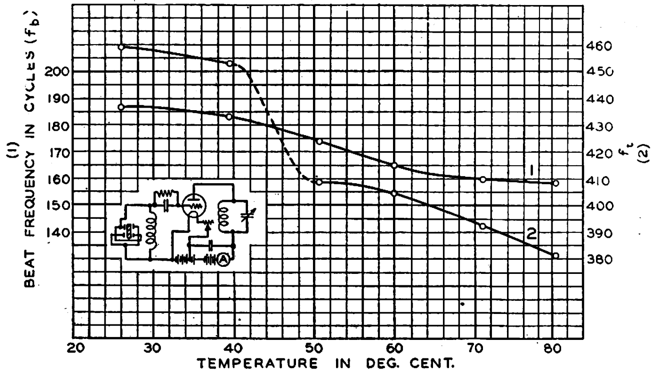

Fig. 14—Excitation of flexural and torsional vibrations in piezo oscillator. Dependence of frequency on temperature. Curve 1, flexural; curve 2, torsional.

4. Quartz Oscillator

There are two modes of cutting a crystal in common use, in one of which the dimension along the normal x is small compared with y, while in the other y is small compared with x, where x is the length along the electric axis, z along the optical axis and y along the third axis.

A piezo oscillator employing the longitudinal vibration is easily excited by the well-known method as shown in Fig. 13, but to obtain lateral (flexural) and torsional vibrations is somewhat difficult. By the connection as shown in Fig. 14, both flexural and torsional oscillations have been tested, where the grid coil is somewhat loosely coupled to the plate coil and four electrodes covering a portion of the surface of quartz are used.[8] When the frequency of the tube oscillator approaches the natural frequency of the quartz, there is in the torsional case a remarkable change in the reading of the ammeter in the plate circuit, by means of which the synchronization can be clearly recognized.

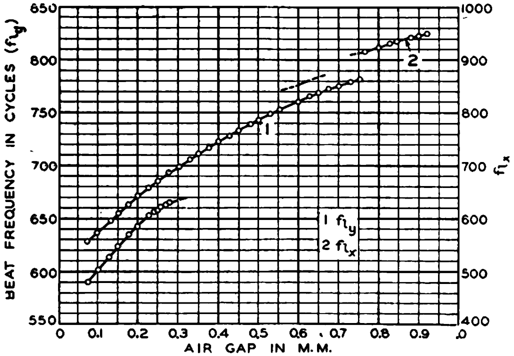

Fig. 15—Dependence of frequency upon air-gap, Curie cut crystal.

The important factors affecting the frequency are (1) the temperature, (2) the air gap in the mounting, and (3) the electric circuit.

The effects of temperature and of the air gap are obtained by the beatnote method as shown in Fig. 17. The temperature coefficient is found to lie between 4.10^{-5} and 1.10^{-5}.

As shown in Fig. 15, the frequency for the Curie cut increases with increase of air gap, and it will be noted that f_{lx} (longitudinal vibration parallel to x-axis) cannot be obtained at some particular length of air gap. This peculiarity disappears on using a mesh electrode instead of the plane electrode.

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[8] The flexural vibrations were of the type described by Harrison, Proc. I. R. E., 16, 1455; November, 1928. The torsional vibrations were as described by Giebe and Scheibe *Zeitschr, f. Phys.* 46 p. 638, 1928.

It is found that the electric circuit does not exert so great an influence as the above-mentioned effects.

IV. RESONATORS

1. Theory of Damping of Resonators

Various types of resonator can serve as efficient working standards of frequency.

A resonator must fulfil the following conditions:

(a) the frequency must depend as little as possible on external conditions, such as the temperature.

(b) the resonant sharpness must be as large as possible.

Resonators may be divided into two classes, one being the electrical type and the other the mechanical type. The frequency bridge and the wavemeter belong to the former, while the sonometer type resonator and the quartz resonator belong to the latter. Now the resonant sharpness Λ can be defined by the following equation:

$$\Lambda = \omega_0 / 2\delta \tag{13}$$

where ω_0 is $2\pi f_0$, f_0 the resonant frequency, and δ the damping factor. Now let Δ, S, R be logarithmic decrement, selectivity, and resonance range, respectively, then we have

$$\Delta = \pi / \Lambda \tag{14)a}$$

$$S = \sqrt{1 + 4.10^4 \Lambda} \tag{14)b}$$

$$R = \frac{\sqrt{3}}{2} \frac{100}{\omega} \% \tag{14)c}$$

The last equation holds when Λ is greater than 100. The numerical relations between these different measures of sharpness of resonance are shown in Table III.

TABLE III

Λ	1	50	100	300	600	1000	3000	6000	10000
Δ	314	6.3×10^{-2}	3.1×10^{-2}	11×10^{-2}	5.2×10^{-3}	3.1×10^{-3}	11.10^{-4}	5.2×10^{-4}	3.1×10^{-4}
S	200	1400	2000	3500	4900	6300	11000	15000	20000
R%	–	–	0.87	0.29	0.15	0.087	0.029	0.015	0.009

It will be clearly understood from the above figures how convenient Λ is to express the sharpness of resonance.

We will next consider the resonant sharpness of the resonator.

2. The Electrical Resonator

The electrical type of resonator may be classified into the wavemeter type and the bridge type.

Campbell's frequency bridge and the ordinary wavemeter illustrate these two types.

Let I be the current in an L-C-R circuit, and I_0 that at resonance, then we have

$$I = \frac{I_0}{1 + j\Lambda\left(x - \frac{1}{x}\right)}. \tag{15}$$

$$I_0 = E/R, \quad x = \omega/\omega_0, \quad \Lambda = 1/\omega_0 CR = \omega_0 L\ R.$$

If the resistance R is that of the inductance coil, Λ is equal to $\tau\omega_0$, where τ denotes the time-constant of the coil. Therefore, the time constant of the coil must be as large as possible, in order to get the greatest sharpness, but since there is a limit to the attainable time constant of a coil, it follows that this resonator, that is, the wavemeter, will be available only for radio frequency. From the above equation, we have

$$\Lambda = \frac{\sqrt{\left(\frac{I_0}{I}\right)^2 - 1}}{x - \frac{1}{x}}. \tag{16}$$

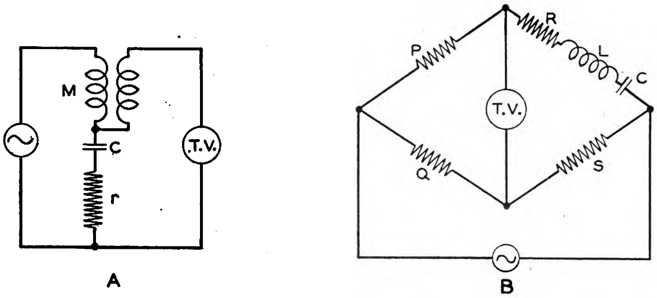

Fig. 16

By this equation, we can obtain the resonant sharpness. In Campbell's bridge as shown in Fig. 16A, let E be the potential difference in the detector circuit and E_0 that at resonance; then supposing that the current at the primary side I is constant, we have

$$E=E_0\left\{1+j\Lambda\left(x-\frac{1}{x}\right)\right\} \tag{17}$$

where $E_0=rI$ and $\Lambda=\omega_0 M/r$.

The resistance r represents the effects of "impurity" of the mutual inductance M (lack of quadrature between primary current and secondary e.m.f.), and dielectric loss in the condenser. If low loss apparatus is used, the sharpness will be very large at audio frequency. From the above equation, it follows that

$$\Lambda=\frac{\sqrt{\left(\frac{E}{E_0}\right)^2-1}}{x-\frac{1}{x}}. \tag{18}$$

By this equation, we can find the resonant sharpness using a tube voltmeter.

With the bridge type resonator as shown in Fig. 16B, supposing that the applied voltage E_0 is always kept constant, we have

$$L=\frac{E}{E_0}\left(1+\frac{S}{Q}\right)=\frac{j\Lambda\left(x-\frac{1}{x}\right)}{1+j\Lambda\left(x-\frac{1}{x}\right)} \tag{19}$$

where the symbol L is introduced as an abbreviation, E is the potential difference in the detector circuit measured by a tube voltmeter T. V., and

$$\Lambda=\Lambda_0\frac{R}{P+R}.$$

Λ_0 denotes the intrinsic resonant sharpness of the resonance arm. From this equation, it follows that

$$\Lambda=\frac{L}{\sqrt{1-L^2}}\Big/\left(x-\frac{1}{x}\right), \tag{20}$$

so that Λ can be found experimentally.

3. The Mechanical Resonator

There are two kinds to consider, the electromagnetic and the electrostatic type. The fork type frequency meter and the magnetic sonometer type resonator belong to the former, while the static sonometer type resonator and the quartz resonator belong to the latter.

In the case of the electromagnetic type resonator, the equation of motion can be written

$$(\alpha D^2+\beta D+\gamma)y=\psi I. \tag{21}$$

The intrinsic sharpness, Λ_0, is given by

$$\Lambda_0=\sqrt{\alpha\gamma}/\beta. \tag{22}$$

This resonator can be represented by an equivalent electric circuit as shown in Fig. 17A, where

$$C_v=\alpha/\psi^2, \quad L_v=\psi^2/\gamma. \qquad R_v=\psi^2/\beta.$$

It will be noticed that the working resonant sharpness, which involved the entire circuit, is different from the intrinsic sharpness. For the electrostatic type of mechanical resonator, the equation of motion can be written

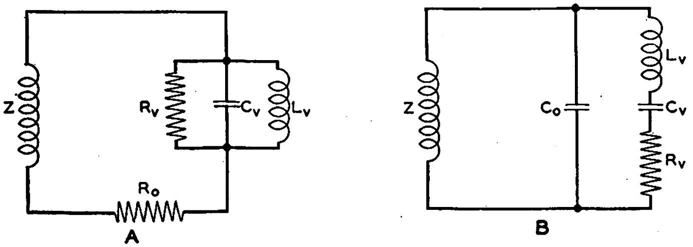

Fig. 17

$$\{\alpha D^2+\beta D+\gamma\}y=\phi E. \tag{23}$$

The intrinsic sharpness is also given by the equation (22).

This resonator can be represented by an equivalent electric circuit as shown in Fig. 17B, where

$$C_v=\phi^2/\gamma \quad L_v=\alpha/\phi^2 \quad R_v=\beta/\phi^2.$$

The working sharpness of the quartz resonator (Figs. 18 and 19) can be written

$$L=\frac{I}{I_0}=\frac{1}{1+j\Lambda\left(x-\frac{1}{x}\right)} \tag{24}$$

and

$$\Lambda = \Lambda_0 \frac{R}{R+R_v}. \tag{25}$$

The shunt capacity C_0 is here assumed to be extremely small.

4. Experimental Results

There are two methods of measuring resonant sharpness. One is the optical method, in which the decay of motion of the vibrator is automatically recorded, and the other is the resonance-curve method using a tube voltmeter or a thermoammeter.

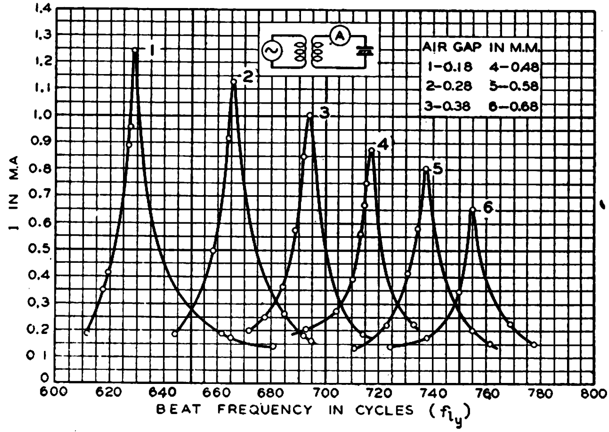

Fig. 18

The author used chiefly the latter method in these experiments. A new type tube voltmeter as shown in Fig. 20 was employed. The author's voltmeter has the following merits: (1) the calibration de-

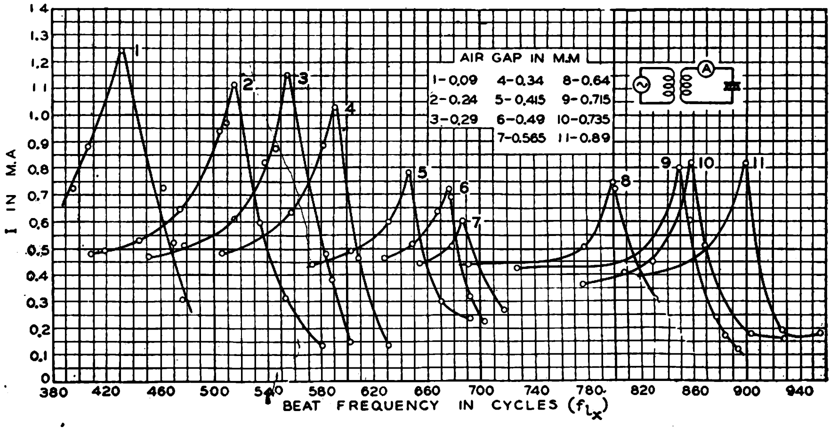

Fig. 19

pends only slightly on the frequency, (2) the input power is extremely small, and (3) the calibration is not affected by the outside circuit.

We can accurately measure a voltage as small as from 0.1 to 1 volt by this voltmeter. The thermoammeter used in these experiments consists of a Western Electric thermovacuum junction and a Weston millivoltmeter. The sonometer type resonator is very useful as a working standard of frequency.

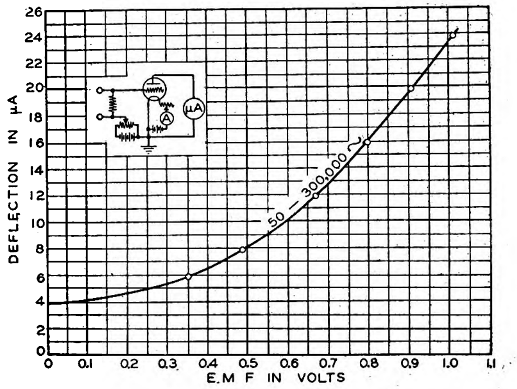

Fig. 20

TABLE IV

Kinds of Resonators		Resonant Frequency	Λ
Hartmann Kempf Frequency Meter		22.5 32.0 46.5 52.5	1100 900 900 610
Tuning Fork		260 490 760	2800 7300 3300
A Freq. Meter Made in Japan		920	6.3
Campbell Frequency Meter		850 640	6.3 2.6
Campbell's Bridge		950 570	11 75
L C Bridge		970 550	17 11
Wavemeter		127000 570000	54 69
G. R. Co. Wavemeter		140000	19
Sonometer	Magnetic	310 730 1010	360–660 630–1400 1000–1600
	Static	990	1600–2000
Piezo Resonator		128000 570000	18000–22000 22000–34000

Many resonance curves have been obtained by the method above mentioned, and the values of resonant sharpness are assembled in Table IV.

The resonant sharpness of the fork type frequency meter and of the tuning fork is measured by the optical method. It will be noted that the resonant sharpness is extremely small with some frequency bridges, and is greatest with the quartz resonator. There are some aperiodic types of frequency bridge, and in such types the calibration curve will be linear, but it will be a drawback that the sharpness of resonance is so small. According to the results shown in Figs. 18 and 19 we see that the resonant sharpness of the quartz resonator decreases regularly with increase of air-gap for the frequency of f_{ly}, while for the frequency f_{lx} the decrease is not regular. It will also be observed that resonance for the frequency of f_{lx} cannot be obtained at some particular length of air-gap and that near this peculiar state the resonant sharpness is very small.

This peculiar phenomenon is considered to arise from the damping in the region of air between the quartz and the electrode.

In conclusion, the author wishes to render thanks to R. Kojima and K. Inagaki, for their very skilful assistance in carrying out the precise measurements.

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FURTHER OBSERVATIONS OF RADIO TRANSMISSION AND THE HEIGHTS OF THE KENNELLY-HEAVISIDE LAYER*

By

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Summary—*The results of further observations on radio transmission phenomena associated with the reflections of radio pulse and spark signals are outlined and a brief theoretical consideration of the form of index of refraction variation best adapted to explain the observed phenomena given. The discussion considers the relation of this index of refraction variation to that discussed in a previous paper and that recently considered by Breit. The results of long-wave field strength observations are also presented. Evidence for a considerable diurnal layer movement is found from the short-wave observations of layer height.*

In a previous paper[1] measurements of the height of the Kennelly-Heaviside Layer as determined from observations made on October 7 and 8, 1928, were presented and the relations between "real" and "virtual" heights as obtained from such observations were discussed.

It is the purpose of this paper to outline and compare the results of similar observations conducted at the Moore School since this paper was submitted for publication and also to present certain results of low-frequency field strength observations (on 18.35 kc) and echo effects on 390 kc.

1. Recent Kennelly-Heaviside Layer Height Determinations on 4435 kc.

(*a*) *Test of Feb. 3 and 4.*

Through the courtesy of the Department of Terrestrial Magnetism and Naval Research Laboratory another 24-hour test was carried out on Feb. 3 and 4 on 4435 and 8870 kc. As in the case of the test of October 7 and 8, the higher frequency was not oscillographable at Philadelphia. Abnormal transmission phenomena were, moreover, observable during this test on 4435 kc to a more marked degree than during the October test. These phenomena indicate what is believed to be somewhat abnormal conditions associated with magnetic storm disturbances prevalent at periods near to but not coincident with this period of observation. As in the case of the previous test, the signal intensity mounted during the sunset period and continued strong for a few hours during the early evening; meanwhile, the layer mounted to a height of 350 km and signals remained strong but were subject to increasing fading. During the late evening and night periods, however, the fading grew progressively worse, while the average signal strength progressively decreased until signals were hardly oscillographable. The oscillographic records during this period showed broadening and blurring of the pulses. These effects accentuated until multiple peaks were no longer observable with such intensity as to render accurate height estimates possible. The results of such esti-

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* Dewey decimal classification: R113.4. Original manuscript received by the Institute, August 21, 1929. Presented before meeting of Philadelphia section, May 24, 1929.

[1] Kenrick and Jen, "Measurements of the Height of the Kennelly-Heaviside Layer," Proc. I. R. E., 17, 711; April, 1929.

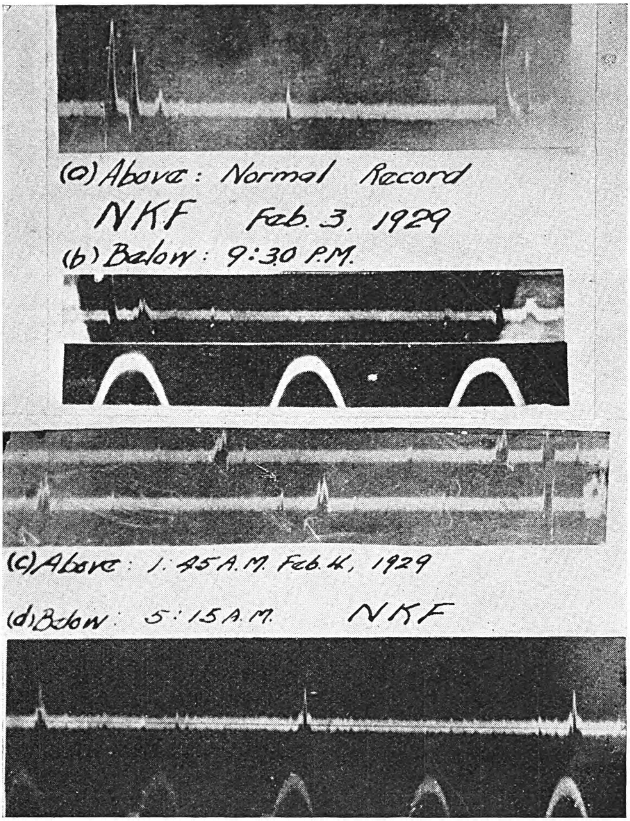

Fig. 1—Oscillographic records of NKF taken on Feb. 3 and 4, 1929.
(a) regular transmission
(b) blurred peaks
(c) increased blurring and fading
(d) single peaks with slight blurring

mates as could be made from faint peaks occasionally observed, however, indicate large virtual heights during the middle night period (accompanied with very weak and broadened peaks). These uncertain virtual height determinations are indicated by the dotted parts of the Feb. 3–4 curve in Fig. 2.

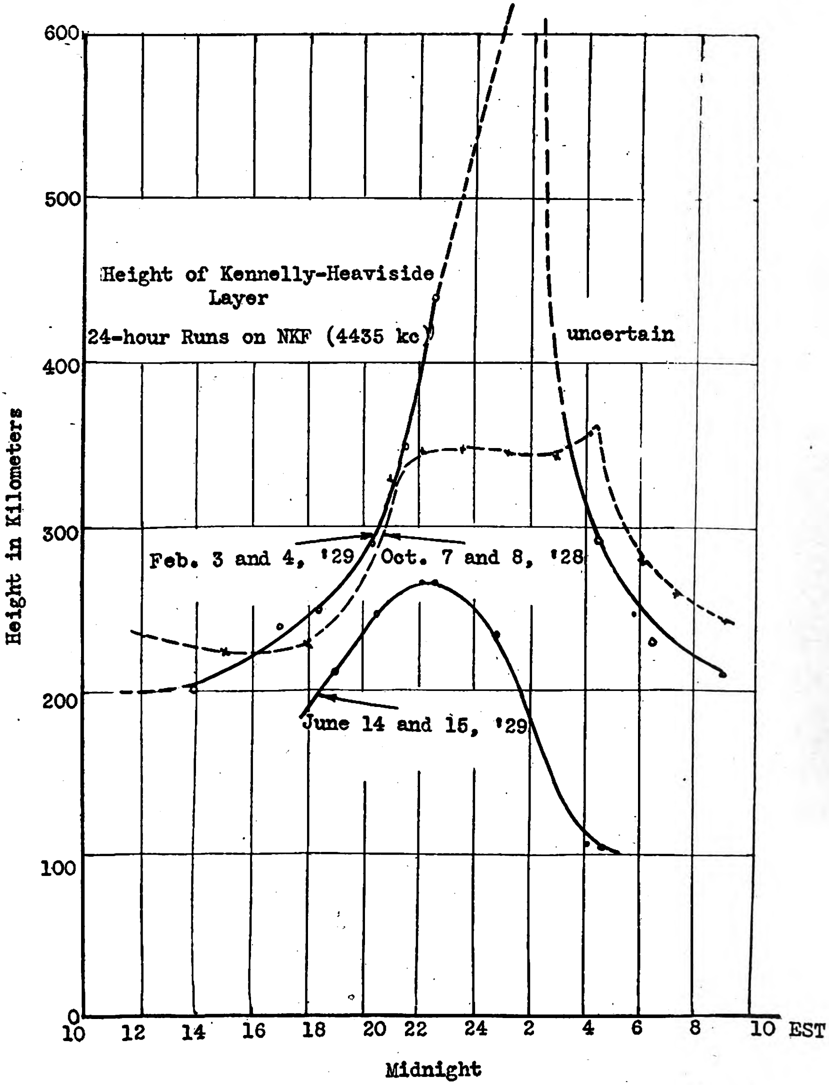

Fig. 2—Comparison of virtual-height data on NKF transmissions as observed at Philadelphia.

Perhaps the most remarkable phenomenon associated with this transmission was observed from 7:30 to 9:00 A.M. (after the sunrise period). In Fig. 3 records taken over this period are shown. It will be noted that the 7:30 A.M. and the 7:55 A.M. records show a second (abnormal) group retarded by nearly but not quite half the group frequency. The 9:15 A.M. record shows sharp and almost single pulses with definite normal group frequency. The existence of long retardations of the form 5000+10000n km (n=0,1,2,3,—) on the relatively low frequency 4435 kc has not to the writers' knowledge been previously observed, although they are of common occurrence on the higher frequencies.[2] The existence of the secondary groups nearly half way between the normal groups suggests a possible doubling of emitted group frequency at the transmitter, but such an explanation was rendered improbable by the failure of the records taken at the Dept. of Terrestrial Magnetism to exhibit such a phenomenon during this period and by the failure of the secondary pulses to be exactly

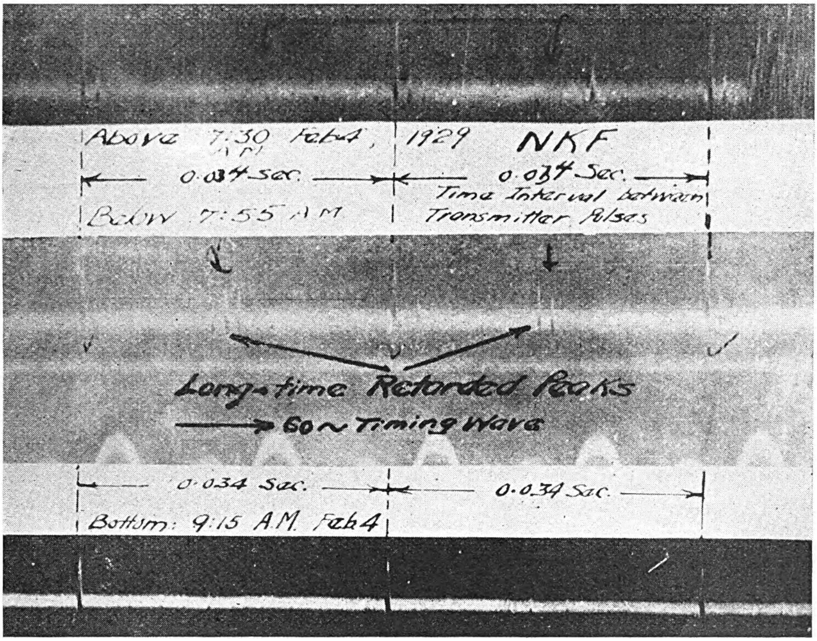

Fig. 3—Long-time retarded peaks of NKF observed at the University of Pennsylvania on Feb. 4, 1929.

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[2] Taylor and Young, "Studies of Echo Signals," Proc. I. R. E., 17, 1491; September, 1929.

centered between the other groups within the tolerance ordinarily observed for the normal pulse groups. Further (less marked) evidence for this phenomenon was also observed on other records taken during the 24-hour interval. The 10000*n* ambiguity is introduced by the normal group frequency together with uncertainty that the secondary group is due to the preceding normal pulse.

(*b*) *Test of June 14–15.*

A similar 24-hour test was again conducted on June 14 and 15. The test was scheduled to begin at 10:00 A.M., E.S.T., June 14, but no audible signals either on 500-cycle modulation or pulses from the multivibrator were received at Philadelphia before 4:00 P.M. After 4:00 P.M. NKF was heard faintly at first, then signals gradually built up to good intensity. The fact that this change took place without any alteration on the setup gives further evidence of high summer daytime attenuation on 4435 kc.

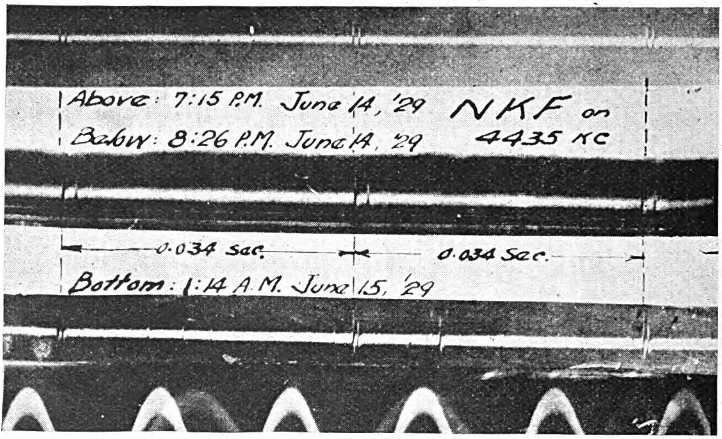

Fig. 4—Oscillograms on NKF transmission taken on June 14 and 15, 1929, showing very low heights.

Referring to Fig. 4, oscillograms taken around 7:00 P.M. show clear-cut peaks. The height of the layer, as determined at this period, is about 200 km (see Fig. 2). An hour later the spacing between the reflected peaks is apparently broadened, showing a larger height. The layer kept increasing in height to about 250 km a little before midnight. After midnight, however, the layer appears to drop down gradually and reaches a level below 100 km at the end of about 4 hours, as is exemplified by the very small spacing between the reflections given by the bottom trace of Fig. 4. In fact the record taken around 4:30 A.M.

June 15 shows almost single peaks. Moreover, the intensity was noticeably faint at this time and became fainter as time went on. Around 7:00 A.M. the signals from NKF were presumably suffering again from as large attenuation as to be quite inaudible at Philadelphia.

Aside from the abnormality of the strangely low layer height after midnight, it will be noticed from Fig. 2 that the entire curve of the height of the layer for this test is lower than on the two preceding tests. This effect is presumably due to summer conditions. It will be noted that the period of this test was nearly coincident with that of longest days and correspondingly high electron density in the upper atmosphere.

2. Observations of Long-Wave Field Strengths.

Considerable evidence has been deduced by numerous competent observers of long-wave field strengths to support the hypothesis that many observed variations are due to phase opposition and reinforcement phenomena.[3]

The observed variations are, however, so complicated as to render an explanation by this hypothesis alone hardly satisfactory. It is evident that a satisfactory explanation of the variations of field intensities for the long waves cannot be realized if we restrict our attention solely to the effects of reflections. Many investigators in the past have already realized the variability of the intensity of radio waves due to absorption, which results from collisions between the electrons and molecules in the upper atmosphere. In fact, the theory of variable attenuation due to changes in ionization offers a beautiful qualitative explanation for the sunrise and sunset phenomena. In general, however, it is difficult to single out the relative importance of the various factors responsible for the change of field intensities. The best we can do, for the present at least, is to find correlations between observations of transmission phenomena on varied frequencies and over varied transmission paths.

Dr. Austin at Bureau of Standards, Washington, D.C., and Mr. Pickard at Newton Centre, Mass., have taken simultaneous recordings of the field intensities of the radio station WCI at Tuckerton, N. J., for a considerable period. The variations of the field intensities at both places have been found to follow in general the same curves.[4]

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[3] Hollingworth, *Jour.* I.E.E., **64**, 579–595; 1926.

Austin, "Experiments in Recording Radio Signal Intensity," PROC. I. R. E., **17**, 1192; July, 1929.

[4] Mr. Pickard kindly informed the authors that this condition was true of 1928, and so far 1929 records show considerable difference between Washington, D. C., and Newton Centre. He also pointed out that perhaps the different Layer height and density this year account for this change.

associated with the relatively brief period of simultaneous observations available are perhaps worthy of note at this time.

Referring to Fig. 5 it is seen that the upper plot shows that the field intensities as recorded at Newton Centre, Mass., and Philadelphia

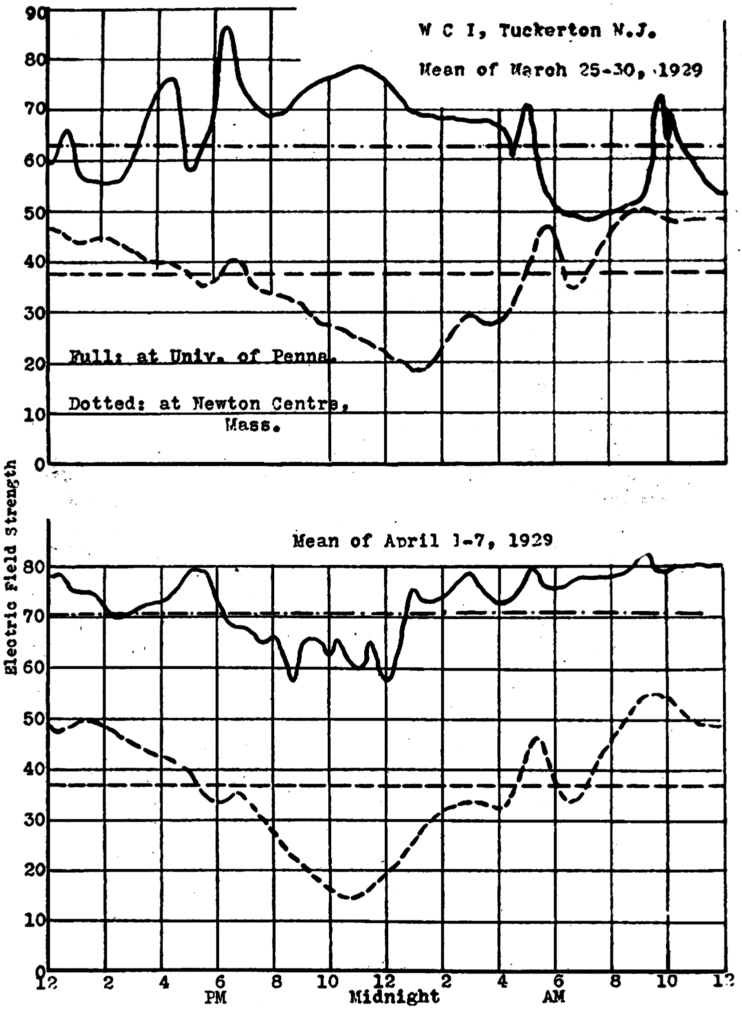

Fig. 6—A weekly mean of the variation of the field strength of WCI.

are just reversed. An extremely low intensity at Newton Centre around midnight corresponds roughly to the very high intensity recorded at Philadelphia. An examination of the daily records during that week shows that there is a gradual shift from the upper plot to the lower plot of Fig. 5. The complete change was made roughly within one week. The lower record shows evidently that the variations of field intensities at two places are just in phase.

Fig. 6 shows the same phenomenon when the averages of the field intensities of the corresponding two weeks are taken. It is seen that although the small variations have been considerably smoothed out the general effect is still clearly illustrated. The existence of such a cyclic change is further evidenced by the reappearance of the opposite patterns for the variations of the field intensities at both places during the third week.

While the period of observation is not long enough to render a general conclusion justified, the above observation of a cyclic change of field intensity suggests a possible correlation with some major changes in the upper atmosphere. It is our hope to confirm the result by further experimentation before making definite conclusions.

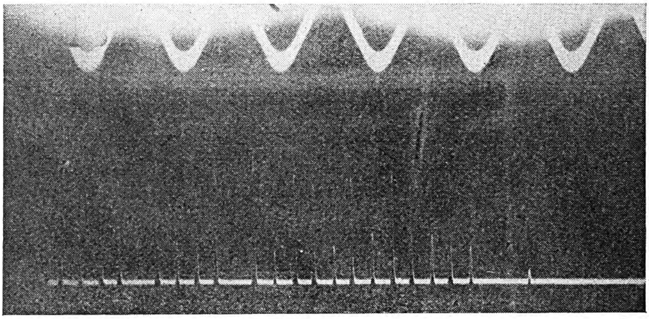

Fig. 7—A typical oscillogram on WNW transmissions as received at the University of Pennsylvania showing single peaks—(variable size due to non-synchronous gap).

3. Observations of Echo Signal Phenomena on 390 kc.

On December 16, 1928, interesting phenomena were observed in a series of simultaneous oscillograms taken at Washington and Philadelphia on transmission from the spark station WNW located at South Wharves, Philadelphia. This station employs synchronous and non-synchronous rotary gaps producing spark discharges of the form shown in Fig. 7.

Simultaneous observations taken at the Moore School of Electrical Engineering during the course of the tests described and involving over 100 oscillograms failed to disclose departures from the form shown in

Fig. 7, except for minor peaks of less than 10 per cent the amplitude of the main peaks; these minor peaks are attributable to the reflected waves of importance in the Washington records but inconsiderable compared to the large low-frequency ground wave at distances of a few miles (i.e. at the Moore School).

During sunrise and sunset periods, however, oscillograms taken at the Department of Terrestrial Magnetism and Bureau of Standards disclose marked doubling of peaks. This phenomenon (by its consistency, progressiveness, and failure to appear in Philadelphia oscillograms taken over the same period) does not appear to be due to any irregularities in the transmitter but rather to a genuine transmission phenomenon.

The phenomena exhibited are of particular interest in exhibiting not only the existence of multiple transmission paths on 390 kc, as would indeed be suspected, but also the importance of absorption at this frequency. It will be noted that by far the most important multiple path phenomena were evidenced during the dawn period, and daytime and evening records failed to show important secondary peaks. The major peak is at this frequency and distance supposedly due to the ground wave.

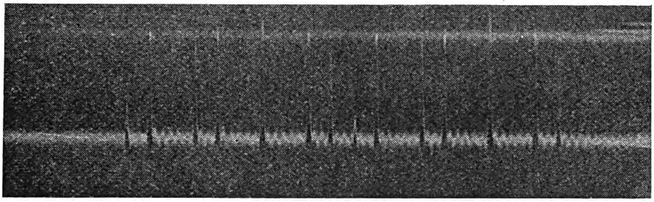

Fig. 8—Spark discharges from WNW being followed, on certain occasions, by minor peaks as received at the University of Pennsylvania.

Virtual height determinations may be carried out from these oscillograms by methods similar to those used with the 4435-kc records and indicate heights of the order of 100 km in accord with the results of Hollingworth and other investigators determining heights by other methods. Considerable evidence is also to be found of a cyclical change in height. Determinations of height from these records are, however, considered somewhat unreliable due to the shortness of the time intervals which must be measured on the records and the failure of the multiple peaks to resolve completely. This renders measurements subject to error because of shifting of the apparent maxima of the peaks as a result of inertial effects and possible phase interferences between waves arriving by various paths (provided the wave trains thus arriving really overlap and the apparent lack of resolution is not due to oscillographic inertia as seems quite possible from theoretical estimates of the duration of the spark discharges based on rational assumption of circuit decrement).

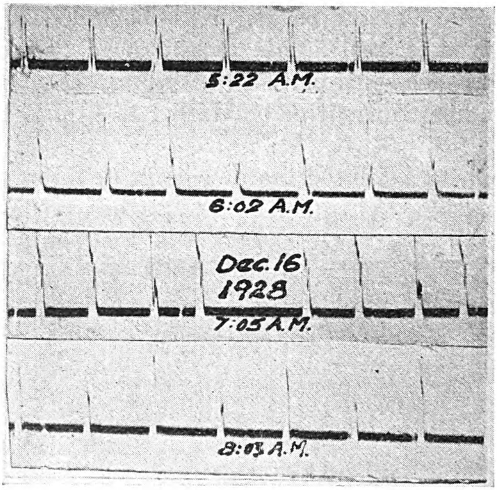

Fig. 9—WNW transmission as received at the Department of Terrestrial Magnetism, Washington, D. C.

4. Approximation Curves for *n* as a Function of Maximum Electric Density and Distribution.

In a previous paper[5] the relations of the virtual heights obtained from pulse records to the maximum height attained by the ray during

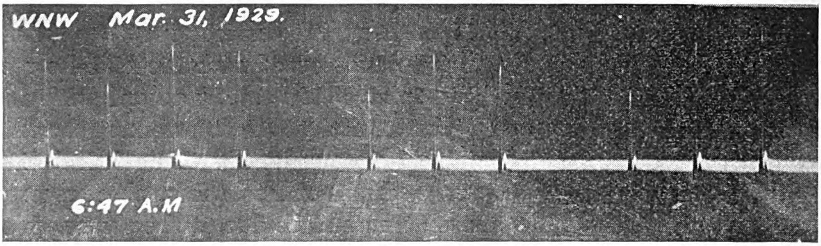

Fig. 10—WNW transmission as received at Bureau of Standards, Washington, D. C.

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[5] Kenrick and Jen, loc. cit.

its trajectory (termed the "true height") were discussed for several types of assumed variation of refractive index with height. One of these assumed types of distributions is compared with the experimentally estimated curve for 4435 kc in Fig. 11.

It will be noted that these curves are concave downward and intersect the *Y* axis nearly parallel to the *n* axis. An examination of the form of the curves for electron density against height reveals that this is a normal condition for frequencies well below the limiting frequency corresponding to that for a skip distance approaching zero.

As the maximum electron density approaches that necessary to turn back a ray at normal incidence, however, the *n* curve assumes a form having a point of inflection such as shown in curve *a* (see Fig. 12).

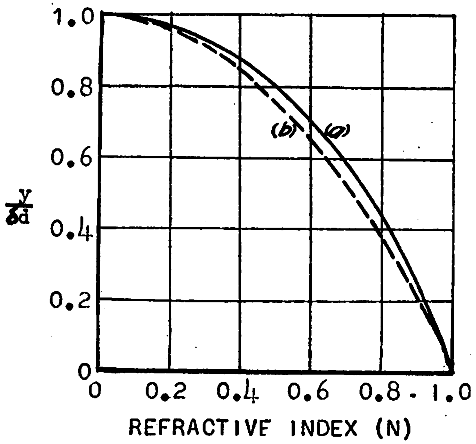

Fig. 11—(a) Experimentally estimated curve at 4435 kc; (b) parabolic curve for the variation of refractive index against height ($n^2=1-ry$).

Further decreases of electron density cause the curve to pass through tangency and finally to assume a shape such as shown in curve *b* corresponding to a finite skip distance. These curves are not well approximated for small values of *n* by the form of equations assumed in the previous paper. We will term the frequency corresponding to the tangent curve the "critical frequency" (obviously variable with the electron density.) Inasmuch as the evidence of the February test indicates phenomena similar to those we would expect near the "critical frequency," a brief resurvey of the problem seems desirable.

5. Approximation Equations for n near the Critical Frequency.

In a recent paper[6] Breit has discussed the problem of long time retardations corresponding to the echo phenomena of Störmer and van der Pol[7] and shown that an exponential distribution of refractive index of the form $n=\epsilon^{-y/a}$ would lead, with the proper choice of constants, to phenomena such as those observed.

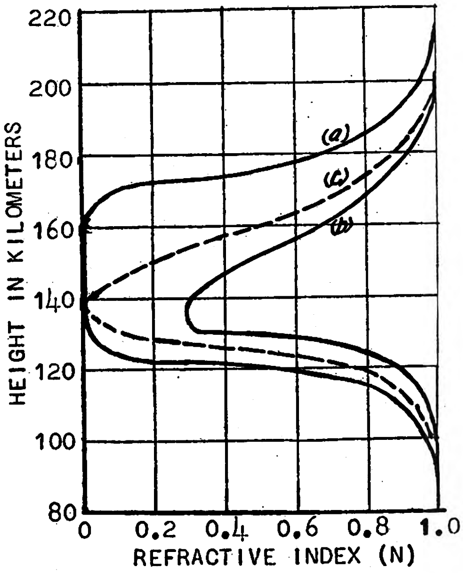

Fig. 12—Variation of refractive index with height
(a) estimated curve at 4435 kc
(b) estimated curve at 8870 kc
(c) estimated curve at 7800 kc (critical frequency)

In Fig. 13 the fits to the n curve of the type c of Fig. 12 by the use of the exponential types of equation and the type leading to a parabolic path are shown.

It will be noted that the exponential type is well adapted to approximate the critical frequency curve for small values of n, while the curves previously considered are satisfactory for larger values of n. In interpreting the phenomena described herein the writers have found a hybrid form in which the exponential type is used for small values of n and the types previously employed for larger values to be successful in describing at least qualitatively the major phenomena observed. In a subsequent section, certain characteristics of pulse echo groups produced by these types of distributions will be discussed and the characteristics of the hybrid distribution then outlined.

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[6] Breit, "Group-Velocity and Long Retardations of Radio Echoes," Proc. I. R. E., 17, 1508; September, 1929.

[7] *Nature*, 122, 878–879, 1928.

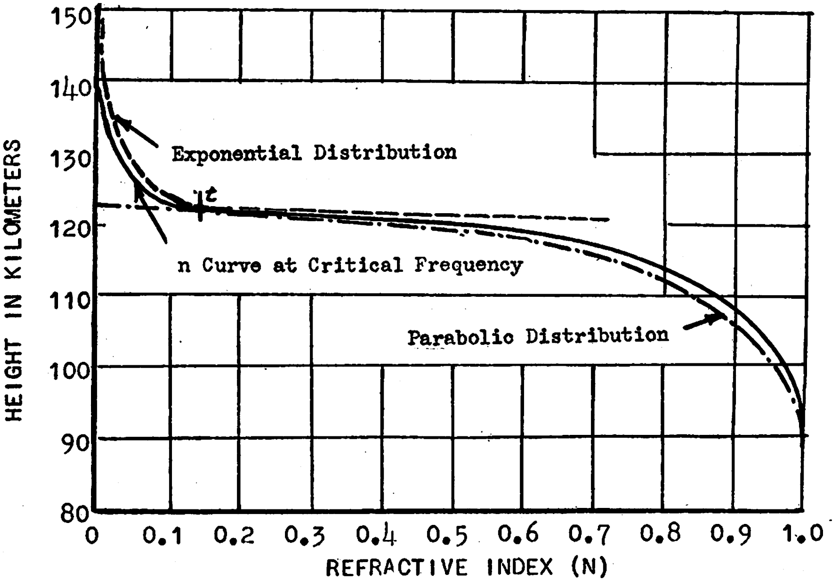

Fig. 13—A comparison between the hybrid distribution for refractive index and the estimated refractive index curve at critical frequency.

6. Virtual Height Determinations in the Absence of a Ground Wave.

Observations made on NKF transmission on 4435 kc and higher frequencies at Philadelphia have not shown evidence of a ground wave, a situation quite in accord with what would be expected from theoretical considerations. Again in accord with theory for this frequency, the Dept. of Terrestrial Magnetism observes both ground waves and sky waves at a distance of 10 km from the transmitter. However, if observations are to be made at frequencies much above the critical (such, for instance, as 8870 kc where the ground wave does not cover up the skip distance) a consideration of what inferences may be drawn as to virtual heights from observations involving sky waves alone seems appropriate. This study is conveniently divided according to the types of dielectric constant distribution encountered.

(a) *Distributions Similar to* $n^2 = 1 - ry$

In cases such as investigated in the previous paper (which may be thought of as more or less normal distributions for frequencies below the critical frequency) it was shown that the ratio of true to virtual height was nearly if not quite constant, and that the virtual height was hence nearly if not quite independent of the angle of incidence of the ray. Under these conditions, if the first two pulses received are assumed to be the first and second reflections the virtual height may be readily evaluated by simple geometry. In fact, for a base line as short as Philadelphia to Washington and virtual heights of the order of 250 km, little error is introduced by taking the observed time interval as twice for a ray to traverse a distance equal to the virtual height (i.e., neglecting the base). The relations involved are clarified by a study of the quantitative results for an illustrative typical case for the $n^2 = 1 - ry$ distribution (see Table I). From Table I it is clear that the intervals between successive pulses are nearly equal to each other (ground wave absent) and to the time for a wave to travel a distance $2h'$ with the velocity of light.

TABLE I

PARABOLIC PATH

$d = 140$ km $(\delta d_1) = 70$ km $\rho = 200$ km $(h/h') \doteq 0.75$

Interval	Time in Sec.	Virtual Height Neglecting Base in km	Virtual Height Considering Base in km	True Height in km
$T_g - T_0 =$	6.67×10^{-4}			
$T_1 - T_g =$	13.2×10^{-4}	197	$h_1' = 280$	$h_1 = 210$
$T_2 - T_1 =$	19.7×10^{-4}	295	$h_2' = 293$	$h_2 = 217$
$T_3 - T_2 =$	19.8×10^{-4}	297	$h_3' = 296$	$h_3 = 218$
$T_4 - T_3 =$	20×10^{-4}	300	$h_4' = 297$	$h_4 = 219$

(b) *Exponential Distribution*

In Table II similar computations have been tabulated for the case of the exponential distribution of n. A lower and more sharply defined boundary was necessarily assumed in this case in order to show several

TABLE II

BREIT'S $\epsilon^{-y/a}$ DISTRIBUTION

$d_0 = 140$ km $h^1 =$ variable $a = 10$ $\rho = 200$ km

Interval	Time in Sec.	Virtual Height Neglecting Base in km	Virtual Height Considering Base in km	True Height in km	$\frac{h}{h'}$
$T_0 - T_g$	6.67×10^{-4}				
$T_1 - T_g$	5.5×10^{-4}	83	154	146	0.95
$T_2 - T_1$	13.0×10^{-4}	195	190	154	0.81
$T_3 - T_2$	23.7×10^{-4}	370	245	160	0.65
$T_4 - T_3$	43.5×10^{-4}	650	345	166	0.48

reflections. It will be recalled that Breit has shown that for this distribution the X coordinate corresponding to the maximum Y ordinate for a ray departing with an angle ϕ_0 in a medium where $n = \epsilon^{-y/a}$ is[8]

$$x = a(\pi/2 - \phi_0) \tag{1}$$

Where n = dielectric constant ϕ_0 = angle of departure.

This function varies but little with ϕ_0 and hence as pointed out by Breit, for a comparable with the distance of transmission ρ, but one value of ϕ_0 is in general possible, although if $2a = \rho$ and $a = d_0$ (the height at which the value of n starts to vary) focussing action will result. For small values of a compared with ρ, however, several reflections may still be possible, but in this case widely variable spacings and virtual heights will be obtained depending on what reflection or path difference is measured.

In general, then, we would expect to find focussing action broad peaks and variable spacing between peaks accompanying distributions of this type. When these phenomena are present, therefore, great caution should be exercised if inferences as to true heights or unique values of virtual heights are to be deduced by measurements of time intervals between received pulses.

(c) *Hybrid Distributions*

Reference to Fig. 13 suggests that a considerably improved fit to the curve we believe to correspond to the critical frequency could be obtained if an equation of the type discussed in paragraph (a) were used for large values of n and an equation of the type of paragraph (b) modified to the form $n = c\epsilon^{-y/a}$ for small values of n, thus providing a point of inflection. The range in a medium approximated by this law for n is readily computable for the sine of the angle of incidence at the point of change is given by the Snell law, i.e.

$$n_t \sin \phi_t = \sin \phi_0 \tag{2}$$

When n_t = dielectric constant at transition point

ϕ_0 = initial angle
ϕ_t = angle at point of transition.

It will be noted that for oblique rays, i.e., $\sin \phi_0 = n_t$ the trajectory of the ray is still normally parabolic (for the $n = \sqrt{1\text{-}ry}$ case).

Appropriate modification of the limits of integration in Breit's equations (by use of equation $V - 1_a$ of the previous paper) gives the values of x and y of the maximum point in the ray path referred to the coordinates of the initial point on the trajectory where n starts to vary, i.e.

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[8] In this paper we are preserving the notation of our previous paper (somewhat different from Dr. Breit's).

$$X = a\left[\frac{\pi}{2} - \sin^{-1}\left(\frac{\sin\phi_0}{n_t}\right)\right] + \frac{2\sin\phi_0}{r}\left[\cos\phi_0 - \sqrt{\cos^2\phi_0 - (1-n_t^2)}\right]$$

or (for small ϕ's)

$$X = a\left(\frac{\pi}{2} - \frac{\phi_0}{n_t}\right) + \frac{2\sin\phi_0}{r}\left[\cos\phi_0 - \sqrt{\cos^2\phi_0 - (1-n_t^2)}\right] \tag{3}$$

$$Y = \frac{1-n_t^2}{r} + a\log\left(\frac{n_t}{\sin\phi_0}\right) \qquad n_t > \sin\phi_0 \tag{4}$$

If preferred, for small values of ϕ_t and $\phi_{0\,1}$, we may write

$$X \doteq a\left(\frac{\pi}{2} - \phi_t\right) + \frac{2\sin\phi_0}{r}\left[\cos\phi_0 - \sqrt{\cos^2\phi_0 - \left[1 - \left(\frac{\phi_0}{\phi_t}\right)^2\right]}\right] \tag{5}$$

$$Y \doteq \frac{1 - \left(\frac{\phi_0}{\phi_t}\right)^2}{r} + a\log(\phi_t^{-1}). \tag{6}$$

It will be noted that for comparable values of a and r, the second terms of the right-hand members of equations (4) and (5) are nearly negligible. A rapid transition hence takes place for values of $\phi \simeq \phi_t$ and for smaller values of ϕ the focussing action described by Breit takes place.

The virtual height for the s'th reflected ray may be readily computed by application of equations (2) to (6). Thus, for the s'th order reflection and a transmission distance ρ, the virtual height for the s'th reflection is (by Breit and Tuve's theorem)

$$h' = \frac{\rho}{s}\tan\phi_s. \tag{7}$$

Where ϕ_s is given (for values of $\phi_s < \phi_t$) by the solution of the transcendental equation

$$\left(\frac{\rho}{s} - d_0\tan\phi_s\right) = a\left(\frac{\pi}{2} - \phi_s\right) + \frac{2\sin\phi_0}{r}\left[\cos\phi_0 - \sqrt{\cos^2\phi_0 - \left[1 - \left(\frac{\phi_0}{\phi_t}\right)^2\right]}\right] \tag{8}$$

Where d_0 = height above earth at which n starts to vary.

ϕ_s = angle of departure from earth of s'th order reflection.

The value of ϕ_s is best derived from transcendental equation (8) by successive approximations. When ϕ_s is greater than $\sin \phi_t$ the first term of the right-hand member should be omitted. When $\sin \phi_t = 1$ equations (7) and (8) may be used to deduce the results given in Table II while the results of Table I are deducible by letting $\sin \phi_t = 0$ (and neglecting the first term of the right-hand member).

The characteristics of the hybrid distributions are, moreover, well adapted to describe the type of phenomena observed during the February test.

Thus, it will be noted that the early night period of the Feb. 3–4 test showed high virtual heights and broadened peaks such as would be given by equations (7) and (8) for a small and $\phi_1 > \phi_t > \phi_2$, i.e., when the critical angle was between the angles corresponding to the first and second reflected paths.

Perhaps more difficult of explanation are the long retardations of the late sunrise period although these are consistent with the hypotheses of an angle $\phi_t < \phi_4$ for part of the path and a value of a approximately $200/\pi$ for part of the path. Under these conditions the first few rays would be normally reflected, giving pulses of normal group frequency while higher order rays after a single normal reflection near Washington would attain a point where $\phi_0 < \phi_t$ on a subsequent entry into the layer. Two successive rays of this type would give rise to a second group of two closely spaced highly retarded rays.

The use of a law for n as a function of heights possessing a compound curvature and an exponential tangency to the axis of y, such as just described, thus appears to possess considerable advantages in the discussion of phenomena such as those considered here. It is at the same time in good accord with the forms of distributions anticipated from other considerations. We see that most of the phenomena described by an exponential distribution may be equally well accounted for by a small region of exponential tangency near the Y axis while normal more oblique reflections phenomena are not impaired. This appears also to be a point worthy of consideration in the case of van der Pol echoes where the long retardation is measured between signals coming through a more or less normal path and a path of high retardation. The presence of both paths is of course essential to the measurement of the retardations.

A study of distributions of this form indicates frequencies near the critical to be the most fertile field for the study of long retardations and similar abnormal transmission phenomena while, as would be suspected, the border line of the skip distance appears the most fertile field in the short-wave region.

CONCLUSIONS

The results of these further Kennelly-Heaviside layer pulse studies indicate that a decided diurnal variation is probably present on 4435 kc. On the whole, the fall test reported in the previous paper was the most satisfactory in furnishing consistent results at Philadelphia and Washington.

The February 3 and 4 test represents a winter run corresponding to a period of short days and low electron densities. The phenomena associated with this run indicate 4435 kc to be close to if not above the "critical" frequency for zero skip distance at periods during the night and dawn at this season and furnish interesting evidence of long time retardations and blurred pulse groups.

The test of June 14–15 indicates a summer condition of low layer height, high electron density and absorption, and probably rather variable ratios of virtual to true heights. The curve shown giving the results of the observations at Philadelphia for this period illustrates these phenomena but also somewhat erratic heights doubtless associated with changes in h/h', etc. This curve represents a good mean of Philadelphia observations but should not be considered of too great intrinsic significance, for Washington observers report erratic and even lower heights (in some cases as low as 70 km) during this test.

The phenomena described on the lower frequency spark signals and the fading records for long waves should all eventually contribute to the large store of data which must be amassed before anything like a complete picture describing the observed phenomena can be evolved.

The highly variable results of the above tests serve to indicate further the desirability of extended consistent observations on variable frequencies and under varied conditions.

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THE CALCULATION OF THE INDUCTANCE OF SINGLE-LAYER COILS AND SPIRALS WOUND WITH WIRE OF LARGE CROSS SECTION*

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1. Introduction

Summary—*Formulas are given for both wires of round cross section and wires of rectangular cross section. These are obtained by an extension of the Rosa Method.*

Tables are given from which the geometric mean distances of rectangles which enter in the formulas may be readily obtained.

THE use of geometric mean distances for the calculation of inductance was first proposed by Maxwell,[1] who obtained the expressions for the geometric mean distances in certain important cases. Formulas for rectangles were developed by Gray,[2] and these were extended to the special cases of parallel and oblique squares by Rosa,[3] who made use of these formulas in the calculation of the correction for insulating space in circular coils of rectangular cross section.

The method of geometric mean distance was also employed by Rosa[4] to obtain the difference between the inductance of a single layer coil of round wire and that of a cylindrical current sheet of the same number of turns, mean radius, and pitch of winding. For this important practical case he gave tables of constants to aid in numerical computations.

The writer has shown[5] by the use of the exact formula of Snow for the inductance of a helix of wire, that the higher order terms, neglected in the simple geometric mean distance method of Rosa, are negligible in practical cases, and that the method is of general use in other important forms of coil.

In the present paper are given tables of the geometric mean distances of equal parallel rectangles, together with working formulas for the calculation of the inductance of windings of round or rectangu-

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[1] Maxwell, Elect. and Mag. II, sects. 691 and 692.

[2] Gray, Absolute Measurements, vol. II, part 1, pp. 288–306.

[3] Rosa, Bull. Bur. of Standards 3, p. 1; 1907.

[4] Rosa, Bull. Bur. of Standards 2, p. 161; 1906. Sci. Paper B. S. no. 169 p. 122, formula (80), Tables VII and VIII, pp. 197–9.

[5] Grover, B. S. Jour. of Research.

lar wire of relatively large cross section. It is hoped to treat the derivation of these formulas and to describe the methods used in calculating the tables in a future paper.

2. Definitions

The geometric mean distance of two areas is that distance whose logarithm is the average of the logarithms of the distances of all the points of one area from all the points of the other. In what follows, if R represents the geometric mean distance, and d the distance between the centers of the areas, we may write

$$\begin{gathered} R = kd \\ \log R = \log d + \log k. \end{gathered} \tag{1}$$

Likewise, the geometrical mean distance of a single area is that distance whose logarithm is equal to the average of the logarithms of the distances of all the points of the cross section from each other. The conception of geometric mean distance is of importance because the mutual inductance of two conductors is found by summing the expression for the mutual inductance of two filaments over the section of the conductors, and since the expression for the mutual inductance involves the logarithm of the distance of the filaments, the geometric mean distance is obtained in the averaging process. Thus a table of geometric mean distances has a use for inductance calculations analogous to that of a table of integrals for mathematical work in general.

3. Nomenclature

Let

b and c be the sides of a rectangular area
ρ = the radius of cross section of a round wire
a = mean radius of a coil or current sheet
n = the number of turns
g = the pitch of a winding
d = the distance between the centers of the cross sections of two turns
r_w = geometric mean distance of the points of the cross section of a wire
r_s = g.m.d. of the cross section of a turn of a current sheet
R_w = g.m.d. of the cross sections of two turns of wire
R_s = g.m.d. of the cross sections of two turns of the current sheet

All dimensions are in centimeters, except where they occur in ratios, in which case it is only necessary that they both be expressed in the same unit. All logarithms, unless otherwise stated, are natural.

Inductances are expressed in millimicrohenries, (μmh), in the formulas; 1 millimicrohenry $= 10^{-9}$ henry. Only steady or low-frequency values of inductance are here considered.

4. Geometric Mean Distances

GEOMETRIC MEAN DISTANCE OF A RECTANGLE

Equation (1) becomes

$$\begin{gathered}\log r = \log (b+c) - 1.5 + \log \epsilon \\ r = k(b+c), \quad \log k = -1.5 + \log \epsilon.\end{gathered} \tag{2}$$

Values of k and $\log \epsilon$ may be obtained from Table I as a function of the ratio b/c (or c/b), whichever is less than unity. Special cases are the straight line for which $\log r = \log b - 3/2$, $r = 0.22313\ b$, and the square for which $\log r = \log b - 0.80508$, $r = 0.44705\ b$.

GEOMETRIC MEAN DISTANCE OF A CIRCLE

(This is given for completeness)

$$\begin{gathered}\log r = \log \rho - \tfrac{1}{4} \\ r = 0.7788\rho.\end{gathered} \tag{3}$$

GEOMETRIC MEAN DISTANCE OF TWO EQUAL PARALLEL RECTANGLES

For convenience of tabulation two cases are considered.

Case 1. Rectangles with their longer sides in the same straight line. (See Fig. 1.) Values of $\log k$ in equation (1) are given for this case in Table II for the two parameters $\gamma = c/d$ and $\delta = b/c$, both of which have values which range from zero to unity. Special cases are straight lines lying in the same straight line with each other $\delta = 0$, and squares $\delta = 1$.

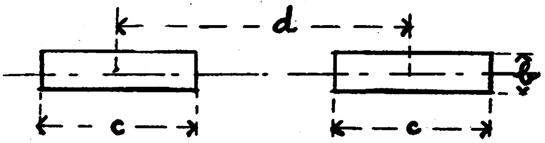

Fig. 1

For rectangles in contact $\gamma = 1$. The values of $\log k$ for squares are positive; the other tabulated values are negative.

Case 2. Rectangles with their longer dimensions perpendicular to the line joining their centers. (See Fig. 2.)

Here it is convenient to employ as parameters $\Delta = c/b = 1/\delta$ and $\beta = b/d$, (or $1/\beta$, whichever is less than unity). An important case is

TABLE I

VALUES OF CONSTANTS FOR THE GEOMETRIC MEAN DISTANCE OF A RECTANGLE

b/c or c/b	k	$\log \epsilon$
0.	0.22313	0.
0.025	0.22333	0.00089
0.05	0.22346	0.00146
0.10	0.22360	0.00210
0.15	0.22366	0.00239
0.20	0.22369	0.00249
0.25	0.22369	0.00249
0.30	0.22368	0.00244
0.35	0.22366	0.00236
0.40	0.22364	0.00228
0.45	0.22362	0.00219
0.50	0.22360	0.00211
0.55	0.22358	0.00203
0.60	0.22357	0.00197
0.65	0.22356	0.00192
0.70	0.22355	0.00187
0.75	0.22354	0.00184
0.80	0.22353	0.00181
0.85	0.22353	0.00179
0.90	0.22353	0.00178
0.95	0.223525	0.00177
1.00	0.223525	0.00177

where Δ is nearly zero, and for those cases $\log k$ changes slowly for small values of Δ. For $\Delta = 0$, the parameter γ used in Table II is zero whatever the dimension b, so that the parameter β is more convenient for this case than γ. Special cases are $\Delta = 1$, squares; $\Delta = 0$, parallel straight lines; and $1/\beta = \Delta$, rectangles in contact. Values of $\log k$ for case 2 are given in Table III. All values are positive.

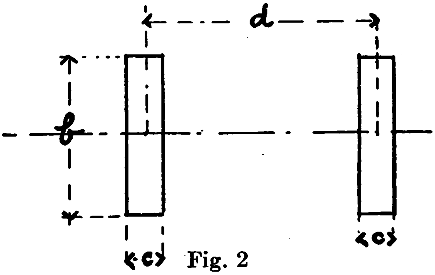

Fig. 2

FORMULAS OBTAINED BY ROSA'S METHOD

Inductance of Single-Layer Coil of Round Wire.

For a coil of round wire of radius ρ, wound with n turns of mean radius a, with a pitch of g, the inductance is given by

$$L = L_s - 4\pi n a(A + B) \qquad (4)$$

in which L_s is the inductance of a cylindrical current sheet of n turns of mean radius a, having the same pitch g, so that its axial length is ng. This may be calculated by known formulas.[6] For an accuracy of a part in a thousand the tables of the writer[7] will be found convenient. The

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[6] Grover, B. S. Jour. of Research, paper 16, October, 1928.

[7] Grover, PROC. I. R. E., 12, 193; April, 1924, p. 193; vol. 11.

TABLE II

VALUES OF THE GEOMETRIC MEAN DISTANCE OF EQUAL PARALLEL RECTANGLES WITH THEIR LONGER SIDES IN THE SAME STRAIGHT LINE

The values of log k in the table are all negative, except in those in the last column, $\delta=1$, (squares).

$\gamma=c/\alpha$	$\delta=0$	0.1	0.2	0.3	0.4	0.5	0.6	0.7	0.8	0.9	$\delta=1.$
0	0	0	0	0	0	0	0	0	0	0	0
	—	—	—	—	—	—	—	—	—	—	+
0.05	0.0002	0.0002	0.0002	0.0002	0.0002	0.0002	0.0001	0.0001	0.0001	0.0000	0.0000
0.10	0.0008	0.0008	0.0008	0.0008	0.0007	0.0006	0.0005	0.0004	0.0003	0.0002	0.0000
0.15	0.0019	0.0019	0.0018	0.0017	0.0016	0.0014	0.0012	0.0010	0.0006	0.0003	0.0000
0.20	0.0034	0.0033	0.0032	0.0030	0.0028	0.0025	0.0021	0.0017	0.0012	0.0006	0.0000
0.25	0.0053	0.0052	0.0051	0.0048	0.0044	0.0039	0.0034	0.0027	0.0019	0.0010	0.0000
0.30	0.0076	0.0076	0.0073	0.0069	0.0064	0.0057	0.0048	0.0038	0.0027	0.0014	0.0001
0.35	0.0105	0.0104	0.0100	0.0095	0.0087	0.0078	0.0066	0.0052	0.0036	0.0018	0.0001
0.40	0.0138	0.0136	0.0132	0.0125	0.0115	0.0102	0.0086	0.0068	0.0047	0.0024	0.0002
0.45	0.0176	0.0174	0.0169	0.0159	0.0146	0.0130	0.0110	0.0086	0.0059	0.0029	0.0003
0.50	0.0220	0.0217	0.0210	0.0198	0.0182	0.0161	0.0136	0.0106	0.0073	0.0036	0.0005
0.55	0.0269	0.0266	0.0257	0.0243	0.0222	0.0197	0.0164	0.0128	0.0087	0.0042	0.0007
0.60	0.0325	0.0321	0.0310	0.0292	0.0267	0.0235	0.0196	0.0152	0.0103	0.0048	0.0010
0.65	0.0388	0.0383	0.0369	0.0347	0.0316	0.0277	0.0231	0.0178	0.0120	0.0055	0.0014
0.70	0.0458	0.0452	0.0435	0.0408	0.0370	0.0324	0.0269	0.0207	0.0137	0.0062	0.0019
0.75	0.0536	0.0529	0.0509	0.0476	0.0431	0.0375	0.0310	0.0237	0.0156	0.0070	0.0023
0.80	0.0625	0.0616	0.0591	0.0551	0.0497	0.0431	0.0354	0.0269	0.0176	0.0075	0.0031
0.85	0.0725	0.0714	0.0683	0.0634	0.0569	0.0491	0.0401	0.0302	0.0195	0.0081	0.0037
0.90	0.0839	0.0825	0.0786	0.0726	0.0648	0.0555	0.0451	0.0337	0.0216	0.0087	0.0046
0.95	0.0973	0.0954	0.0903	0.0828	0.0734	0.0625	0.0504	0.0374	0.0236	0.0092	0.0056
1.00	0.1137	0.1106	0.1037	0.0942	0.0828	0.0700	0.0561	0.0413	0.0258	0.0098	0.006

TABLE III

VALUES OF THE GEOMETRIC MEAN DISTANCES OF EQUAL PARALLEL RECTANGLES WITH THEIR LONGER SIDES PERPENDICULAR TO THE LINE JOINING CENTERS

The table gives values of log k as a function of $B=b/d$ or $1/\beta=d/b$, whichever is less than unity, and for $\Delta=c/b$. 5 In all cases log k is positive.

β	$\Delta=0$	0.1	0.2	0.3	0.4	0.5	0.6	0.7	0.8	0.9	1.0
0	0	0	0	0	0	0	0	0	0	0	0
0.1	0.0008	0.0008	0.0008	0.0008	0.0007	0.0006	0.0005	0.0004	0.0003	0.0002	0.0000
0.2	0.0033	0.0033	0.0032	0.0030	0.0028	0.0025	0.0021	0.0017	0.0012	0.0007	0.0000
0.3	0.0074	0.0073	0.0071	0.0067	0.0062	0.0056	0.0048	0.0038	0.0027	0.0015	0.0001
0.4	0.0129	0.0128	0.0124	0.0118	0.0109	0.0098	0.0084	0.0068	0.0050	0.0027	0.0003
0.5	0.0199	0.0197	0.0191	0.0182	0.0169	0.0152	0.0131	0.0106	0.0077	0.0043	0.0005
0.6	0.0281	0.0278	0.0271	0.0258	0.0240	0.0216	0.0185	0.0152	0.0111	0.0064	0.0011
0.7	0.0374	0.0371	0.0361	0.0344	0.0320	0.0290	0.0251	0.0206	0.0155	0.0090	0.0019
0.8	0.0477	0.0473	0.0461	0.0440	0.0411	0.0373	0.0321	0.0268	0.0200	0.0129	0.0031
0.9	0.0589	0.0584	0.0569	0.0544	0.0506	0.0464	0.0404	0.0338	0.0254	0.0158	0.0046
1.0	0.0708	0.0702	0.0685	0.0655	0.0614	0.0560	0.0492	0.0406	0.0313	0.0199	0.0065
0.9	0.0847	0.0841	0.0821	0.0787	0.0738	0.0675	0.0596	0.0501	0.0382	0.0250	
0.8	0.1031	0.1023	0.0999	0.0959	0.0903	0.0829	0.0745	0.0622	0.0485		
0.7	0.1277	0.1268	0.1240	0.1192	0.1125	0.1037	0.0925	0.0788			
0.6	0.1618	0.1607	0.1573	0.1507	0.1436	0.1329	0.1194				
0.5	0.2107	0.2094	0.2053	0.1984	0.1886	0.1754					
0.4	0.2843	0.2826	0.2776	0.2691	0.2567						
0.3	0.4024	0.4003	0.3942	0.3831							
0.2	0.6132	0.6105	0.6021								
0.11	1.0787	1.1075									
$1/\beta$	$\Delta=0$	0.1	0.2	0.3	0.4	0.5	0.6	0.7	0.	0.9	1.0

constants A and B were calculated by Rosa[4] and tabulated. The constant B is a function of the number of turns and is reproduced here as Table IV. In this, small errors in the original table have been corrected by the use of an improved method of calculation. Rosa's table of A is not given here. Its calculation by the relation $A=5/4-\log g/\rho$ is readily accomplished using the Table VI for computing natural logarithms from common.

TABLE IV

THE CONSTANT B IN FORMULA (6) FOR COILS OF ROUND WIRE

n	B	n	B	n	B
1	0.	31	0.3087	110	0.3278
2	0.1137	32	0.3095	120	0.3285
3	0.1663	33	0.3102	130	0.3291
4	0.1973	34	0.3109	140	0.3296
5	0.2180	35	0.3115	150	0.3301
6	0.2329	36	0.3121	160	0.3305
7	0.2443	37	0.3127	170	0.3309
8	0.2532	38	0.3132	180	0.3312
9	0.2604	39	0.3137	190	0.3315
10	0.2664	40	0.3142	200	0.3318
11	0.2715	41	0.3147	220	0.3323
12	0.2758	42	0.3152	240	0.3327
13	0.2795	43	0.3156	260	0.3330
14	0.2828	44	0.3160	280	0.3333
15	0.2857	45	0.3164	300	0.3336
16	0.2883	46	0.3168	350	0.3341
17	0.2906	47	0.3172	400	0.3346
18	0.2927	48	0.3175	450	0.3349
19	0.2946	49	0.3179	500	0.3351
20	0.2964	50	0.3182	550	0.3354
21	0.2980	55	0.3197	600	0.3356
22	0.2994	60	0.3210	650	0.3357
23	0.3008	65	0.3221	700	0.3358
24	0.3020	70	0.3230	750	0.3360
25	0.3032	80	0.3246	800	0.3361
26	0.3043	85	0.3253	850	0.3362
27	0.3053	90	0.3259	900	0.3362
28	0.3062	95	0.3264	1000	0.3364
29	0.3071	100	0.3269	∞	0.3379
30	0.3079				

Inductance of a Flat Spiral of Round Wire

The spiral is assumed to consist of n turns of round wire of cross sectional radius ρ, wound with a pitch g. The inductance is calculated by the previous formula, (4), except that here a represents the radius of the mean turn of the spiral and for L_s is put the inductance L_s' of a disk current-sheet. The corrections A and B are the same as in the previous case, and their values may be obtained as has already been indicated.

The inductance of the equivalent current sheet L_s may be obtained in most practical cases by means of the formula[8]

$$L_s' = 4\pi a n^2 \left[\left(\log \frac{8a}{ng} - \frac{1}{2} \right) + \frac{n^2 g^2}{96a^2} \left(\log \frac{8a}{ng} + \frac{43}{12} \right) + \cdots \right]. \qquad (5)$$

However, in the case of spirals wound nearly to the center, this formula is not sufficiently convergent, and Spielrein's formula[9] should be employed.

[8] Rayleigh and Niven, *Proc.* Roy. Soc., **32**, pp. 104–141; 1881. B. S. Sci. Paper 169, p. 117, formula (70).

[9] Spielrein, *Arch. für Electrotech.* **3**, p. 187; 1915. B. S. Sci. Paper 320, p. 555; 1918. Butterworth, *Proc.* Lond. Phys. Soc., **32**, part 1, p. 31; 1919.

Single-Layer Coil of Rectangular Cross Section

The cross section of the wire is assumed to have dimensions b and c, the former being radial. Two different formulas may be used. In the first

$$L = L_s - 4\pi na(A_1 + B_1) \tag{6}$$

L_s being the inductance of the equivalent cylindrical current sheet, calculated as for the single-layer coil of round wire.

$$-A_1 = \log \frac{g}{b+c} - \log \epsilon \tag{7}$$

$$\begin{aligned} B_1 = B + 2\Bigg[& \frac{n-1}{n} \log k + \frac{1}{12}(\beta^2 - \gamma^2)\left\{(S_2 - 1) - \frac{(\log n/2 + 1.270)}{n}\right\} \\ & - \frac{1}{60}\left(\beta^4 + \gamma^4 - \frac{5}{2}\beta^2\gamma^2\right)\left\{(S_4 - 1) - \frac{(S_3 - 1)}{n}\right\} \\ & + \frac{1}{168}\left(\beta^6 - \gamma^6 + \frac{5}{3}\beta^2\gamma^4 - \frac{14}{5}\beta^4\gamma^2\right)\left\{(S_6 - 1) - \frac{(S_5 - 1)}{n}\right\}\Bigg] \end{aligned} \tag{8}$$

in which $\beta = b/g$, $\gamma = c/g$, $\delta = b/c$, and $\Delta = c/b$. The value of $\log k$ is taken from Table II or III, B from Table IV, and $\log \epsilon$ from Table I. The remaining constants are

$$S_2 - 1 = 0.6449 \qquad S_4 - 1 = 0.0823$$

$$S_3 - 1 = 0.2021 \qquad S_5 - 1 = 0.0369$$

$$S_6 - 1 = 0.0173$$

When the dimension b of the cross section, which is perpendicular to the line joining the centers of the turns, is large, it is convenient also to calculate the inductance by the expression

$$L = L_s'' + 4\pi na(A_2 + B_2) \tag{9}$$

(note that the correction for cross section is positive here). In this formula L_s'' is the inductance of a current sheet of n turns having a mean radius a, the current being uniformly distributed over a rectangular cross section whose radial dimension is b, and whose axial dimension is ng. The calculation of L_s'' may be conveniently made by the formulas and tables of the Bureau of Standards Scientific Paper 455.

The correction constant A_2 is given by the formula

$$A_2 = \log \frac{b+g}{b+c} + (\log \epsilon_3 - \log \epsilon_w) \tag{10}$$

in which $\log \epsilon_s$ is the value of $\log \epsilon$ taken from Table I for the ratio b/g or g/b, and $\log \epsilon_w$ is the value for the ratio b/c or c/b.

The remaining correction term is

$$\begin{aligned}B_2=2\Bigg[&\frac{n-1}{n}(\log k_s-\log k_w)\\&-\frac{1}{12}(1-\gamma^2)\left\{(S_2-1)-\frac{(\log n/2+1.270)}{n}\right\}\\&+\frac{1}{24}\left\{\beta^2(1-\gamma^2)-\frac{2}{5}(1-\gamma^4)\right\}\left\{(S_4-1)-\frac{(S_3-1)}{n}\right\}\\&+\frac{1}{36}\left\{\beta^2(1-\gamma^4)-\beta^4(1-\gamma^2)-\frac{3}{14}(1-\gamma^6)\right\}\left\{(S_6-1)-\frac{(S_5-1)}{n}\right\}\Bigg]\end{aligned} \tag{11}$$

$\log k_s$ and $\log k_w$ are to be taken from Table III for the parameters $1/\beta=g/b$ and $\Delta=g/b$ for the former and g/b and c/b for the latter. As before $\gamma=c/g$, and the S constants are those given above.

Inductance of a Flat Spiral of Rectangular Wire

In this case the dimension b of the rectangular cross section is axial and c is radial. Formulas (6) and (9) are immediately available, with the understanding that L_s' is the inductance of a disk current sheet of mean radius a and radial width ng, while L_s'' is the inductance of the current sheet of mean radius a whose rectangular cross section has a dimension b in the axial direction and ng in the radial. This case is illustrated in example 1 below.

Inductance of Polygonal Single-Layer Coils and Spirals

Putting p for the length of the side of the polygon, (in the case of the spiral, the side of the mean turn), N for the number of sides of the polygon, the formulas for the equivalent current sheets are

$$L_p=2Nn^2p\left[\log\frac{p}{ng}+\left(\frac{3}{2}-r\right)+\frac{s}{3}\,\frac{ng}{p}+\frac{t}{6}\,\frac{n^2g^2}{p^2}\right] \tag{12}$$

for the single-layer coil, and

$$L_p'=2Nn^2p\left[\log\frac{p}{ng}+\left(\frac{3}{2}-r\right)+\frac{s'}{3}\,\frac{ng}{p}+\frac{T}{6}\,\frac{n^2g^2}{p^2}\right] \tag{13}$$

for the spiral.

The constants involved in these formulas are given in the following table:

TABLE V

	r	s	s'	t	T
Triangle	1.4055	2.2092	1.2396	−11/12	23/12
Square	0.77401	1	0.5328	−0.0429	3/4
Hexagon	0.15152	0.3954	0.2033	+0.1160	0.2947
Octagon	−0.21198	0.2146	0.1090	+0.1052	0.1662

The corrected formulas for cross section for a winding of round wire are

$$L=L_p-2Npn\left[A+B+\frac{s}{3}\,\frac{g}{p}+\frac{t}{6}n\frac{g^2}{p^2}\right] \tag{14}$$

for the single-layer coil, and

$$L=L_p'-2Npn\left[A+B+\frac{s'}{3}\,\frac{g}{p}+\frac{T}{6}n\frac{g^2}{p^2}\right] \tag{15}$$

for the spiral. The quantities A and B are the same as for circular coils and spirals of round wire. These are multiplied here by a factor which is twice the total length of wire, just as was the case with circular coils and spirals.

For wire of rectangular cross section the only change necessary is to replace A and B by the quantities A_1 and B_1 in formulas (14) and (15).

Long polygonal solenoids are best calculated by finding the inductance of the equivalent circular solenoid.[10]

EXAMPLES

Example 1. A flat spiral has ten turns of wire 0.2 cm in diameter, wound with a pitch of 1 cm. The mean radius of the innermost turn is 20.5 cm, and that of the outermost is 29.5 cm. Thus the mean radius is 25 cm. Also $g=1$, $n=10$, $\rho=0.1$. The inductance of a disk current sheet of ten turns of mean radius of 25 cm and radial width $ng=10$ cm is 78.75 μh. In formula (4) $A=5/4-\log 10=-1.0526$, and from Table IV, for $n=10$, $B=0.2664$. Thus, $(A+B)=-0.7862$, and the correction is $786.2\pi=2.47$ μh. Thus the total inductance of the spiral is $78.75+2.47=81.22$ μh.

If the same spiral were wound with wire of rectangular cross section 1 cm by 0.1 cm, with the smaller dimension radial, then in (6) $L_s'=78.75$ μh as before, $b=1.0$, $c=0.1$, $\Delta=0.1$, $\beta=1$, $\gamma=c/g=0.1$. From Table I for $c/b=0.1$, $\log \epsilon=0.0021$, and from Table IV, $B=0.2664$. Log k is taken from Table III for $\beta=1$ and $\Delta=0.1$, and is found to be 0.0702. Since $\log g/b+c=-\log 1.1=-0.0953$ it is found from formulas (7) and (8) that

$$A_1=0.0953+0.0021=0.0974$$

$$B_1=0.2664+2\Big[0.9(0.0702)+\frac{1}{12}0.99(0.6449-0.2879)$$
$$-\frac{1}{60}0.975(0.0823-0.0202)+\frac{1}{168}0.972(0.0173-0.0037)\Big]$$
$$=0.2664+2[0.06318+0.02945-0.00102+0.00008]=0.4497.$$

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[10] Grover, B. S. Sci. Paper 468, p. 753, and Table 1, p. 781; 1923.

Thus $(A_1+B_1)=0.5471$, and the correction is -1.72 μh. The inductance of the spiral is $78.75-1.72=77.03$μh.

If formula (9) be employed to solve the same problem, we have to find first the inductance of a circular current sheet of mean radius 25 cm, with a rectangular cross section whose axial dimension is 1 cm, and whose radial is $ng=10$ cm. The value is $L_s''=75.70$μh. From Table I, using the ratio $b/g=1$, $\log \epsilon_s=0.00177$, and for $c/b=0.1$, $\log \epsilon_w=0.0021$. $\log (b+g/b+c)=\log (2/1.1)=0.5978$, so that from formula (10) $A_2=0.5975$.

Table III gives for $\beta=1$, $\Delta=1$, $\log k_s=0.0065$, and with $\beta=1$, $\Delta=0.1$, $\log k_w=0.0702$. Also $\gamma=0.1$. Thus from formula (11)

$$\begin{aligned} B_2=2\Big[&0.9(0.0065-0.0702)\\ &-\frac{0.99}{12}(0.6449-0.2879)\\ &+\frac{1}{24}\left\{0.99-\frac{2}{5}0.9999\right\}\{0.0823-0.0202\}\\ &+\frac{1}{36}\left\{0.9999-0.99-\frac{3}{14}\right\}\{0.0173-0.0037\}\Big]\\ =2[&-0.05733-0.02945+0.00153-0.00010]=-0.1707\end{aligned}$$

$(A_2+B_2)=0.4268$. The correction is therefore 1.34μh, and the inductance of the spiral comes out $75.70+1.34=77.04$μh, which checks the other method very closely.

APPENDIX

TABLE VI

FOR CONVERTING COMMON LOGARITHMS INTO NATURAL LOGARITHMS

Common	Natural	Common	Natural	Common	Natural	Common	Natural
0	0.0000	25.0	57.565	50.0	115.129	75.0	172.694
1.0	2.3026	26.0	59.867	51.0	117.432	76.0	174.996
2.0	4.6052	27.0	62.170	52.0	119.734	77.0	177.299
3.0	6.9078	28.0	64.472	53.0	122.037	78.0	179.602
4.0	9.2103	29.0	66.775	54.0	124.340	79.0	181.904
5.0	11.513	30.0	69.078	55.0	126.642	80.0	184.207
6.0	13.816	31.0	71.380	56.0	128.945	81.0	186.509
7.0	16.118	32.0	73.683	57.0	131.247	82.0	188.812
8.0	18.421	33.0	75.985	58.0	133.550	83.0	191.115
9.0	20.723	34.0	78.288	59.0	135.853	84.0	193.417
10.0	23.026	35.0	80.590	60.0	138.155	85.0	195.720
11.0	25.328	36.0	82.893	61.0	140.458	86.0	198.022
12.0	27.631	37.0	85.196	62.0	142.760	87.0	200.325
13.0	29.934	38.0	87.498	63.0	145.063	88.0	202.627
14.0	32.236	39.0	89.801	64.0	147.365	89.0	204.930
15.0	34.539	40.0	92.103	65.0	149.668	90.0	207.233
16.0	36.841	41.0	94.406	66.0	151.971	91.0	209.535
17.0	39.144	42.0	96.709	67.0	154.273	92.0	211.838
18.0	41.447	43.0	99.011	68.0	156.576	93.0	214.140
19.0	43.749	44.0	101.311	69.0	158.878	94.0	216.443
20.0	46.052	45.0	103.616	70.0	161.181	95.0	218.746
21.0	48.354	46.0	105.919	71.0	163.484	96.0	221.048
22.0	50.657	47.0	108.221	72.0	165.786	97.0	223.351
23.0	52.959	48.0	110.524	73.0	168.089	98.0	225.653
24.0	55.262	49.0	112.827	74.0	170.391	99.0	227.956
25.0	57.565	50.0	115.129	75.0	172.694	100.0	230.259

Example 2. To illustrate the use of Table VI, suppose that the natural logarithm of 0.08549 is required. The common logarithm is 2.93192. This may be written as 0.93192−2.

Table VI is a multiplication table constructed for the factor 2.302585 . . . ., the ratio between natural and common logarithms, so that if we denote this ratio by M, then from the table

$$\begin{array}{rr}
0.93\ M= & 2.14140 \\
0.0019\ M= & .00044 \\
0.00002M= & \underline{.00005} \\
 & 2.14189 \\
-2M= & \underline{-4.6052\ } \\
 & -2.4633\ \
\end{array} = \text{required natural logarithm.}$$

* * *

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OPERATING CHARACTERISTICS IN PHOTOELECTRIC TUBES*

By
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Summary—*This paper gives a discussion of the characteristics of photoelectric tubes which are important in the engineering applications of such tubes. Definitions of photometric terms are given together with the introduction of a few special terms essential to photoelectric tube work. Curves are given showing both the static and dynamic characteristics of a typical tube. A simple mathematical analysis is given of the elementary photoelectric tube circuit. An appendix of photometric formula and conversion factors is provided.*

INTRODUCTION

THERE has been a great deal published on the physics of the photoelectric tube, but very little on its engineering aspects. The photoelectric tube now has such wide application that it is becoming of increasing importance to have a system of engineering terminology and to consider its characteristics in detail. The object of this paper is to discuss some of the photoelectric terms and tube characteristics which are of use in connection with the engineering applications of the tubes.

A knowledge of the photometric terms in most general use is essential to an understanding of photoelectric tube operation. For that reason the following list of definitions is given for reference.

ILLUMINATING ENGINEERING NOMENCLATURE AND PHOTOMETRIC STANDARDS[1]

Luminous flux is the rate of passage of radiant energy evaluated by reference to the luminous sensation produced by it.

Lumen. The unit of luminous flux is the lumen. It is equal to the flux emitted in a unit solid angle by a uniform point source of one international candle.[2]

Luminous intensity of a point source in any direction is the luminous flux per unit solid angle emitted by that source in that direction.

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* Dewey decimal classification: 535.3. Original manuscript received by the Institute, July 18, 1929.

[1] *Trans.* I. E. S., July, 1925.

[2] For the calculation of the light flux on a given area the following formula is found convenient:

$$l = C \times (\text{Area})/(\text{distance})^2.$$

l—light flux in lumens. C—the candle power of the source in the direction considered. Area and distance are in corresponding units.

Candle power is luminous intensity expressed in candles.

Illumination at a point of a surface is the density of the luminous flux incident at that point or the quotient of the incident flux by the area of the surface when the latter is uniformly illuminated.

Foot Candle. Taking the foot as the unit of length, the unit of illumination is the lumen per square foot. It is known as the foot candle.

PHOTOELECTRIC TUBE CHARACTERISTICS

From the viewpoint of performance and characteristics there are two distinct types of photoelectric tubes: the vacuum tube and the gas-filled tube. The first, as the name implies, is highly evacuated, while the second contains an inert gas at pressures in the order of 20 to 150 microns. The anode voltage-anode current characteristic of a typical vacuum photoelectric tube with a constant value of light flux incident on the cathode is shown in Fig. 1. This characteristic is similar to thermionic emission when limited by filament temperature. When an inert gas at low pressures is present the characteristic changes to that of the gas tube shown in Fig. 1.

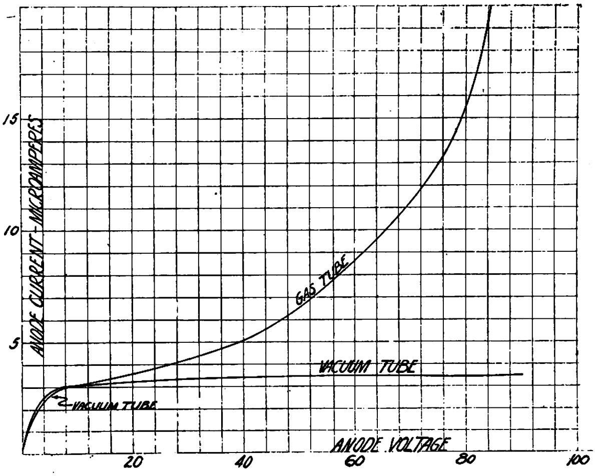

Fig. 1—Anode voltage-anode current characteristic of photoelectric tubes.

If the anode voltage is held constant and the light flux varied, the current is very nearly proportional to the amount of light. A typical family of curves of this type is shown in Fig. 2.

Although the photoelectric emission of electrons from the cathode responds to light variations of radio frequency, the amplification due to the gas does not follow such high-frequency variations. Fig. 3

shows the variation of the a-c output with frequency of the alternating light input for a typical gas-filled photoelectric tube. The lower curve shows the response of a vacuum photoelectric tube under similar conditions.

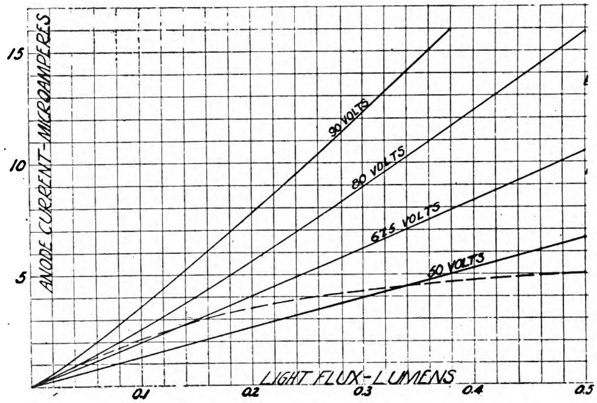

Fig. 2—Light flux-anode current characteristic of gas-filled photoelectric tubes.

Photoelectric Terms

Sensitivity. The slope of the light flux-anode current curve at a given point is the sensitivity of the tube at that point. That is, $S = di_a/dl$. This characteristic of the photoelectric tube is more clearly understood if the light flux is compared to the grid voltage on the triode and the sensitivity to the mutual conductance. It must be noted, however, that the relation of each to change in anode voltage is not similar.

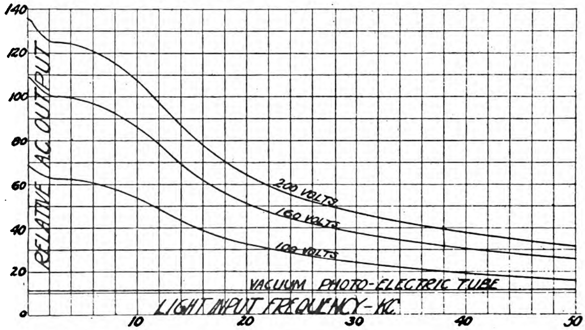

Fig. 3—Dynamic sensitivity of gas-filled photoelectric tubes with 100 per cent modulation of light flux.

Conductance Per Lumen. The slope of the anode voltage-anode current curve divided by the light flux is the conductance per lumen at the point considered. This may be expressed as

$$G = \frac{1}{l} \frac{di_a}{de_a}.$$

This factor determines the maximum voltage output for a given condition of fidelity. This characteristic may be compared to the plate conductance of the triode. It must be borne in mind, however, that the conductance is a direct function of the light flux in the photoelectric tube case.

APPLICATIONS

Before any calculation of circuit constants can be made, it is necessary to consider the applications to which the photoelectric tube is to be placed. The majority of the applications fall into one of three classes.

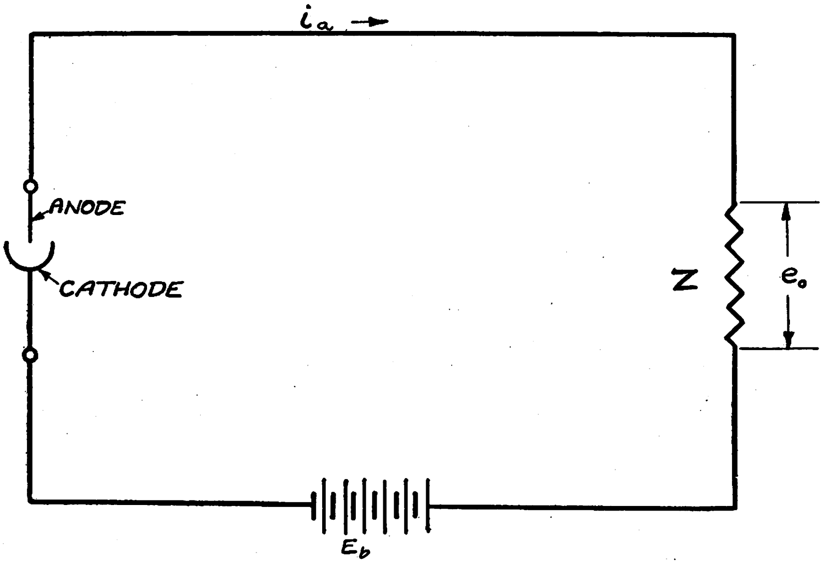

Fig. 4—Elementary circuit for photoelectric tube operation.

The first is d-c relay operation. In this class the tube is used to actuate a relay to either an "on" or "off" position.

The second is d-c linear operation. The photometry of lamps or the recording of daylight intensity is representative of this group.

The third group may be called a-c linear operation. That is, the response must be linear and uniform throughout a definite frequency range. Television and talking motion pictures are examples of such requirements.

Each of the above classes of operation requires an independent set of operating conditions.

CIRCUIT THEORY

In order to determine the best operating conditions, it is necessary to analyze the elementary circuit of the photoelectric tube in detail. The fundamental circuit considered is that shown in Fig. 4.

The following notation will be used in the development of the circuit equation:

l —instantaneous value of light flux on the cathode
i_a —anode current
S —sensitivity
G —conductance per lumen
Z —load circuit impedance
e_o—output voltage

From the figure it is seen that anode current is composed of two components. Let i_{a1} be the component which flows due to the constant battery voltage E_b, and i_{a2} be the component due to the change in anode voltage. Thus,

$$i_a = i_{a1} + i_{a2}.$$

By reference to the definitions of sensitivity and conductance per lumen it is seen that

$$i_{a1} = Sl$$

$$i_{a2} = (lG)e_0$$

then

$$i_a = Sl + lGe_0.$$

However,

$$e_0 = -i_a Z$$

so

$$i_a = Sl - lGi_aZ.$$

Simplifying

$$i_a = \frac{Sl}{1 + lGZ}$$

then

$$e_o = \frac{SlZ}{1 + lGZ}.$$

By comparison with the triode plate current equation,

$$i_p = \frac{e_g G_m}{1 + G_p R_0}$$

it is seen that the fundamental difference is that while the triode equation is linear with respect to grid voltage, the photoelectric tube equation is hyperbolic with respect to the similar controlling factor, light flux. The dotted line in Fig. 2 shows the variation of anode current with light flux when a 5-megohm load resistance is used.

It is the curvature of this equation that limits the output voltage to a definite value for a given quality of reproduction. The magnitude of the harmonics introduced may be calculated from the current equation,

$$i_a = \frac{Sl}{1 + lGZ}$$

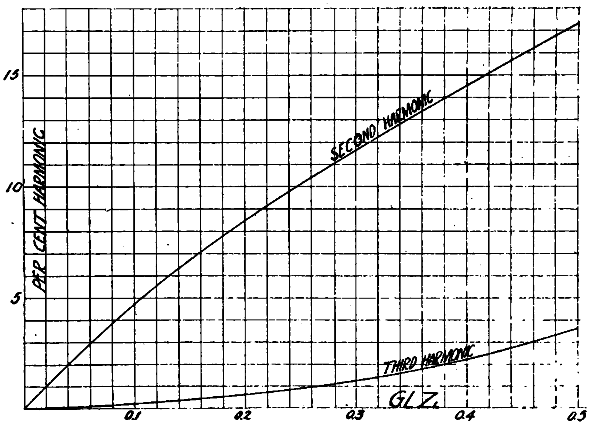

Fig. 5—Harmonics introduced into the output as a function of the product of conductance per lumen, light flux, and load impedance.

by substituting L (sin $\omega t+1$) for l. This assumes 100 per cent modulation of the light flux, but little error is introduced by the steady component.

Substituting and simplifying

$$i_a = \frac{S}{GZ}\left(1 - \frac{1}{GZL(1+\sin \omega t)+1}\right).$$

Expanding into a Fourier series

$$i_a = \frac{S}{GZ}\left(\begin{array}{l} A_0 + A_1 \sin \omega t + A_2 \sin 2\omega t \cdots \\ \quad + B_1 \cos \omega t + B_2 \cos 2\omega t \cdots \end{array}\right).$$

It is now possible to determine the ratio of the harmonic currents to the fundamental, and the magnitude of the fundamental frequency.

These are found to be a function of the product of G, Z, and L. The per cent harmonics are plotted in Fig. 5, while the output voltage is obtained from Fig. 6.

Due to the large value of load circuit impedance necessary to secure maximum output voltage, the capacity shunting the load circuit should always be considered. This capacity often limits the load impedance to a small value when it is required to cover a large frequency band.

In general, the major consideration in the design of photoelectric tube circuits is the fact that the photoelectric tube is fundamentally an exceedingly high impedance device. For that reason the load impedance can never approach the condition for maximum power output, but should always be as large as conditions of fidelity and capacity will allow.

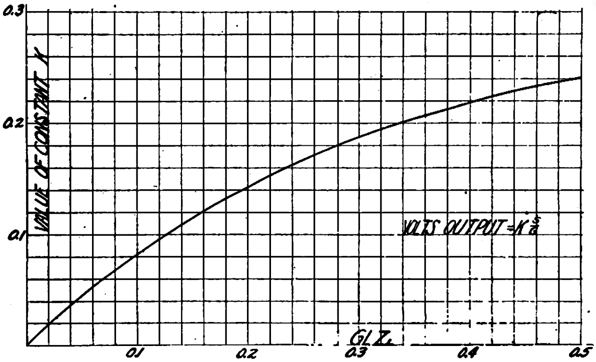

Fig. 6—Output voltage of fundamental frequency as a function of conductance per lumen, light flux, and load impedance.

Appendix

The following list of formulas and conversion factors is given for convenience in future reference.[3]

Photometric quantity	Name of unit	Symbol and defining equation	Abbreviation for name of unit
1. Radiant flux	Erg per sec. Watt	Φ	
2. Luminous flux	Lumen	F	l.

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[3] *Trans.* I. E. S., July, 1925.

3.	Luminous intensity (candlepower)	Candle	$I=\dfrac{dF}{d\omega}$	c.
4.	Illumination	Lux, phot foot-candle	$E=\dfrac{dF}{dS}$ (incidence)	ph., fc.
5.	Quantity of light	Lumen-hour	$Q=\int F dt$	l-hr.
6.	Brightness	Candle per unit area Lambert millilambert Foot-lambert	$B=\dfrac{dI}{dS \cos \theta}$	 L mL. ft.L.
7.	Visibility		$K=\dfrac{F\lambda}{\phi\lambda}$	
8.	Reflection factor		p	
9.	Absorption factor		a	
10.	Transmission factor		τ	
11.	Mean spherical candlepower		scp.	
12.	Mean lower hemispherical candlepower		lcp.	
13.	Mean upper hemispherical candlepower		ucp.	
14.	Mean zonal candlepower		zcp.	
15.	Mean horizontal candlepower		mhc.	

16. A source of unit spherical candlepower emits 12.57 lumens.
17. 1 lumen is emitted by a source whose spherical candlepower is 0.07958.
18. 1 lux = 1 lumen incident per square meter = 0.0001 phot = 0.1 milliphot.
19. 1 phot = 1 lumen incident per square centimeter = 10,000 lux = 1,000 milliphots = 1,000,000 microphots.
20. 1 milliphot = 0.001 phot = 0.929 foot-candle.
21. 1 foot-candle = 1 lumen incident per square foot = 1.076 milliphots = 10.76 lux.
22. 1 lambert = 0.3183 candle per square centimeter = 2.054 candles per square inch = 929 foot-lamberts.
23. 1 millilambert = 0.001 lambert = 0.929 foot-lambert.
24. 1 foot-lambert = 1.076 millilamberts = 0.00221 candles per square inch = 0.000343 candles per square centimeter.
25. 1 candle per square centimeter = 3.1416 lamberts = 2919 foot-lamberts.
26. 1 candle per square inch = 0.487 lambert = 487 millilamberts = 452 foot-lamberts.

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THE PENETRATION OF ROCK BY ELECTROMAGNETIC WAVES AND AUDIO FREQUENCIES*

BY

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Summary—*This paper aims to throw some light upon the much discussed topic of radio penetration and the path of radio waves. Conclusive experimental evidence at the Mammoth Cave of Kentucky, free from all metallic conductors, confirms the hypothesis that radio waves penetrated through the ground and did not follow air openings or conductors. There is further evidence that audio-frequency magnetic fields are not greatly damped through an overburden to 300 ft. of sandstone and limestone having a resistivity of about 50,000 ohms per cm³.*

DURING the month of June, research work has been carried out at the Mammoth Cave, Kentucky, under the joint auspices of the United States Bureau of Mines and the Geological Survey of Canada. The disputed question of the extent to which wireless waves will penetrate into the earth has a high scientific interest, and it bears directly on some practical modern methods of geophysical prospecting for underground conductors such as the sulphides of lead, iron, and copper.

Previous investigations have been made at the experimental mine of the Bureau of Mines at Bruceton, Pa., in the Caribou mine, Colorado, and in the Mount Royal Tunnel, Montreal.[4] Notices and letters on this subject have also appeared from time to time in the columns of *Nature*. The results of all this work have been of a dubious character because of the presence of iron pipes and rails, and of copper conductors connected with electric lighting. There were three schools of thought; some claimed that the radiations came through the entrance, others that the waves passed through the rock, while the remainder considered that the effects were transmitted along wires, pipes, and rails. Probably all three paths are available, but certainly the Mount Royal experiments proved beyond question that the short waves of length 40 m (7,500 kc), failed to enter the tunnel by any of these means for more than a few hundred feet. Longer waves, broadcasting upwards, were heard and measured throughout the entire three and a half miles of tunnel.

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* Dewey decimal classification: R113.5. Original manuscript received by the Institute, July 6, 1929. Published by permission of U. S. Bureau of Mines and the Canadian Geological Survey.

[4] An account of the experiments may be found in Technical Paper 434, pp. 37–40, Bureau of Mines, Department of Commerce, Washington, D. C., and in PROC. I. R. E., 17, 347: Feb., 1929.

To settle the question, Mammoth Cave, Kentucky, was selected for experiments, and it proved a fortunate choice. This cave has not been electrified for illumination, and the managers were good enough to remove telephone wires. There are no conductors whatever in the cave except here and there short lengths of iron handrails. The exits are filled by and sealed with the waters of Echo River and the River Styx, which flow into Green River. A few years ago a new entrance was found and some electric wiring introduced; this is four miles away overland and eight miles away by cave. A careful survey was made both under and above ground so that corresponding points could be located and levels known. Most of the work was carried out above River Hall, with 300 ft. of overburden,—about 200 ft. of Mammoth Cave limestone capped by 100 ft. of Cypress sandstone. The resistivities of these rocks *in situ* were ascertained by electrical prospecting to be of the order of 10,000 to 20,000 ohms per cm^3.

BROADCASTING WAVELENGTH

The words and music from Louisville, Nashville, and Cincinnati were received on a No. 26 portable Radiola superheterodyne six-tube set, and their bearings were taken with the small loop which is part of the apparatus. These signals were traced into the cave from the entrance and were lost at 500 ft. from the mouth, where the overburden is 150 ft. Yet under the Mammoth Dome, 1,000 ft. in the cave, capped by an ascertained thickness of 75 ft., mostly sandstone, the signals were clear enough, and these were traced along a rather narrow tunnel for 300 ft., when the thickness was 150 ft.

On a later occasion Messrs. Joyce, Barlow, and Kidd took a 300-ft. aerial and gave three turns of it round the loop, so as to form a close coupling, and led the other end to the ground. In this manner there was at night good audible broadcasting in River Hall 300 ft. below the ground, and more than 1,000 ft. from the entrance at a point approached by the famous tortuous passage well named the Corkscrew. On the next night, signals and speech were also audible at Echo River, which is reputed to be 350 ft. underground. The conclusion is irresistible: waves from distant stations can be detected under 300 ft. of sandstone and limestone—waves which do not come through the entrance, and which do not pass along conductors.

LONG-WAVE STATIONS

Morse signals from about six stations with bearings approximately N. 65 W. (Long Island and others) were heard on a loop with a model RE low-frequency receiving equipment, consisting of an antenna coupling unit, a radio-frequency amplifier, a low-frequency receiver, and a tuned audio amplifier. These signals were clear enough to hear, but their intensities were hard to measure, partly owing to static. The wave fronts appeared to be mainly vertical, and the waves must therefore have been travelling through rock rather than down from above. The reception was again at River Hall, 300 ft. below the surface plateau near the camping ground of the Mammoth Cave Hotel. The bearings of these signals obtained above and below ground were nearly identical. This picture of a wave, with its front mainly vertical, travelling through the rock (just as through the air), is rather novel. We could pick up these unidentified stations with three turns of wire on a 4-ft. square frame, and we obtained bearings therewith.

HORIZONTAL WAVE FRONTS

A circular horizontal transmitting loop, 100 ft. in diameter, was placed on the ground, and the insulated wire could be tapped so as to give one to ten turns as required. A rectangular receiving coil of three turns (40×10 sq. ft.) was placed on the floor of the cave at River Hall (300 ft. down). Currents of known magnitude of about half an ampere were introduced into a single turn of the loop, using a gasoline engine generator to excite an oscillator. The high-frequency loop current was modulated by applying a 500-cycle a-c potential on the oscillator plates. The frequency was varied by steps of 10, from 20 to 110 kc. The received voltage on the coil was measured in microvolts, and averaged about four, suggesting a flux of 10^{-9} lines per sq. cm in the cave, due to half an ampere in the loop above ground. It was a most remarkable fact that the 20- and 30-kc frequencies boomed out loudly as compared with the 40- and 100-kc frequencies. Measurements made above ground on the same frequencies showed no such emphasis, and there is a strong suggestion that these frequencies (20 to 30 kc) pass with markedly less absorption through the rock. This question must stand over for a fuller investigation. The selectivity did not appear to be a function of the receiving instrument.

AUDIO FREQUENCIES

When 500-cycle frequency alternating current was impressed on the full ten turns of the 100-ft. diameter circular loop, powerful signals were received with headphones both with and without a three-stage amplifier, using a 400-turns 3×2 sq. ft. rectangular coil, at depths of 100 and of 300 ft. Of particular interest is the fact that detection was readily made without amplifier in the Mammoth Dome, 900 ft. on an inclined line from the horizontal loop. We conclude that the electromagnetic effects of a 500-cycle frequency passed through 900 ft. of continuous rock. It is intended to publish a full report on these experiments in due course.

Proceedings of the Institute of Radio Engineers
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A VACUUM-TUBE VOLTAGE REGULATOR FOR LARGE POWER UNITS*

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Summary—*A new voltage regulator is described, suitable for large power units of standard construction. This regulator uses vacuum tubes as the controlling element, in place of vibrating contacts that have hitherto been employed. The voltage variation is less than 0.1 per cent when the load is gradually changed from no load to a little above full load. When this regulator is applied to d-c unit, some a-c source is necessary to supply the B-eliminator which is an essential part of it. The voltage of this a-c source, however, need be only fairly constant; it may safely vary by ±3 volts (in the case of a 110-v source) without causing more than ±0.1 per cent variation in the d-c line under regulation. When an a-c unit is to be regulated, on the other hand, the same unit may supply the B-eliminator.*

ACCURATE regulation of voltage is one of the serious problems encountered in connection with the generation and distribution of electrical power. In research laboratories this problem may may be of prime importance, since even a small variation of voltage may seriously interfere with certain types of work.

There are two main types of regulators now on the market. One of these, the induction regulator, is applicable only to a-c units; the other type consists essentially of a control magnet which, operated by the line voltage, actuates a pair of contacts that in turn control the field of the main unit in some way. This may be accomplished by short-circuiting a resistance in series with the main field or, in case of large machines, by short-circuiting the field of an auxiliary motor, the armature of which is in the main field circuit. Vibrating contacts as control elements have obvious disadvantages. A slight amount of wear or a little dirt may hamper the regulation seriously or may even start an accumulative deterioration. Moreover, the contacts introduce a time lag in addition to that which is inherently present in the main field and in the field of the auxiliary motor, thus delaying the response.

Such a regulator, with a counter e.m.f. auxiliary motor, has been in use in these laboratories for some time. The regulation obtained with it was not, however, sufficiently close for various types of precision work. Furthermore, it was not reliable. At times the line voltage would suddenly drop three or four volts and then oscillate about a lower mean value. The cause of this trouble was later found to be a sudden reversal of the auxiliary motor. With proper adjustments of

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* Dewey decimal classification: R358. Original manuscript received by the Institute, April 24, 1929.

speed the possibility of reversing was minimized. These troubles led to the development of a new regulator, which utilizes vacuum tubes as control elements. The description of this regulator forms the subject matter of this paper.

Various suggestions have been made regarding the application of thermionic tubes to voltage regulation problems. Van der Bijl[1] has suggested two very simple arrangements which are adaptable only to specially designed generators and tubes. During the World War a small generator was developed in conjunction with a special tube for airplane radio work, where a small amount of power was to be delivered at a constant voltage under highly variable speed.[2,3] Recently Stoller and Power have made use of a vacuum-tube amplifier in a precision voltage regulator for small amounts of a-c power.[4] Thus various schemes have already been worked out for special applications. The regulator described here is believed to be the first vacuum-tube regulator that can be applied to any standard d-c machine. With a slight modification it can also be used for a-c machines, as will be pointed out later. Probably the main reason why vacuum tubes have not yet been used for this purpose is that the requirements have never been exacting enough to warrant the expense and maintenance of the apparatus involved. The difficulties of maintenance have been minimized by the employment of a B-eliminator in the new regulator. The operation is very simple, since the control has been centralized in a single rheostat. With the exception of the exciter generator, which requires a special field winding, all the parts used are standard and can be readily procured.

DESCRIPTION

Fig. 1 is a schematic representation of the regulator. It consists of a one-stage high-μ d-c amplifier (UX240) which amplifies the fluctuation of the line and feeds into a power tube (UX210) through a coupling resistance, R_3. The output of the power tube is passed through the field of the exciter generator which supplies the main generator field. The only batteries used are in the input of the first tube, employed to balance out most of the line voltage (in this case 110 volts). Since the grid of this tube has a negative bias, no current passes through these batteries, and their life is at least the normal shelf-life. The regulator is controlled by rheostat R_1 in the input circuit, mounted on the main switchboard. Balancing voltage, for adjusting the power tube grid bias, and all B-power are supplied by the eliminator, which is not shown in the diagram. Filaments are lighted by alternating current supplied by independent transformers.

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[1] "The Thermionic Vacuum Tube and Its Applications," pages 371–373.

[2] E. B. Craft and E. H. Colpitts, "Radio Telephony," *Trans.* A. I. E. E., 38, part 1, page 330, 1919.

[3] Robt. W. King, "Thermionic Vacuum Tubes and Their Applications," *Bell Sys. Tech. Jour.*, 2, No. 4, p. 96–98, Oct. 1928.

[4] H. M. Stoller and J. R. Power, "A Precision Regulator For Alternating Voltage," *Jour.* A. I. E. E., XLVIII, No. 2, p. 110, February, 1929.

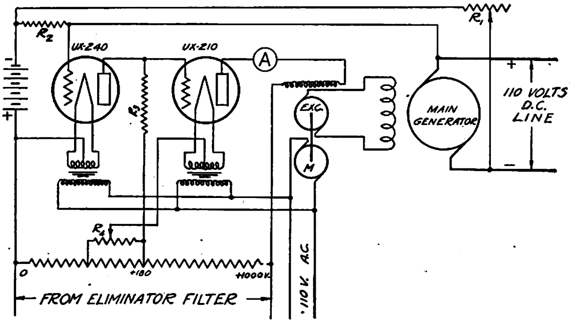

Fig. 1

The field of the exciter generator is designed so that its resistance matches the plate resistance of the power tube and at the same time it operates well below the knee of the saturation curve for the exciter, in order to get the maximum possible change in the output for a given variation of the line voltage.

In Fig. 2 is shown the actual photograph of the instrument with the exciter motor-generator set. The exciter may be run by the same

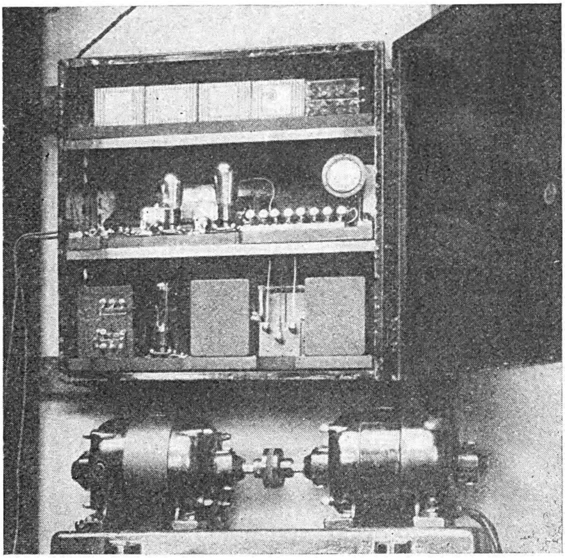

Fig. 2

prime mover as the main generator. It was found more convenient under local conditions to drive it by a separate motor which gets its power from the same line as the B-eliminator through a common transformer connected in parallel with the motor that drives the main generator.

The cycle of operation is as follows. A small decrease of line voltage produces a corresponding increase of negative grid bias on the 240 amplifier tube, thereby lessening its plate current. This reduction of the plate current which flows through the coupling resistance R_3, decreases the drop across it and consequently decreases the negative bias of the 210-power tube. The plate current of the latter tube thus increases and boosts the exciter voltage and in turn the line voltage. This process is reversed throughout when the line voltage tends to rise.

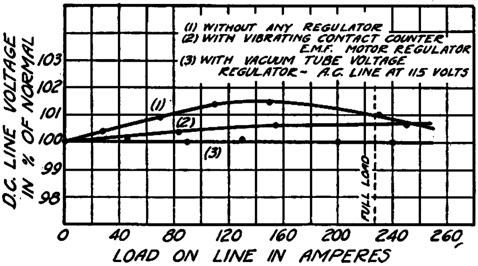

Fig. 3

Performance

The results of various tests are shown in Figs. 3, 4, and 5. Without any regulator the generator does not give a constant voltage at a given load. It drifts continuously. Curve 1 of Fig. 3, therefore, merely shows an estimated mean value of the line voltage without any regulator. Curve 2 shows the performance of a vibrating contact counter-e.m.f. motor type regulator. Curve 3 of the same figure shows that the new regulator maintains the line voltage constant at all loads tested, within the reading error of the voltmeter, which is ±0.1 per cent. In Fig. 4 is shown the regulation as a function of the a-c voltage applied to the eliminator which supplies the amplifier and the power tube. Curve (1) of this figure shows that a 30 per cent variation in a-c voltage makes less than 1 per cent change in d-c voltage. Hence we may safely say that the regulator will function quite satisfactorily on an average a-c line.

Under the direction of Professor Bedell, a transformer has recently been developed in this laboratory that delivers a constant load at a constant voltage over a large range of primary voltage. Such a transformer, designed for this purpose, maintains 115 volts at the secondary terminals when loaded with the eliminator. The primary voltage was changed gradually from 100 to 130 volts without making any readable change in the line voltage, as shown by curve 2 of Fig. 4.

Fig. 5 shows continuous records of line voltage taken under normal operating conditions on a Leeds and Northrup recorder. Each ordinate division corresponds to about 1.2 per cent variation from the mean value of the line voltage. While these records were being taken, the load varied at random, as would be expected in a physical laboratory.

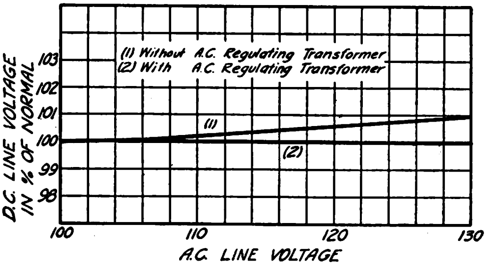

Fig. 4

When record 5 (Fig. 5) was under way, it was observed that, in addition to the general variation, the load changed ±5 amperes continuously at intervals of about one minute. In examining these records, it must be borne in mind that, due to the inductance of the main field, a sudden change in load causes a corresponding jump in the voltage, which returns to normal in the course of a fraction of a second. These variations are too fast for the recorder to follow and only once in a while does one of these dips show up in the records. The time lag caused by the inductance of the main field also sets a limit to the rapidity with which any regulator can compensate for voltage fluctuations. For this reason, regulation for small rapid fluctuations is no better with this regulator than with the vibrating contact type.

The various records shown in Fig. 5 were taken under the following conditions:

No. 1 is the record of line voltage when no regulator was in operation.

No. 2 was taken with the vibrating contact counter-e.m.f. motor type regulator. Voltage, though kept well within reasonable range during this period, fluctuated quite rapidly.

Nos. 3 and 4 are the records taken with the new regulator while the eliminator was directly connected to the a-c line. The marked difference between these two records is caused by the fact that when No. 3 was being taken there was a large air compressor on the a-c line.

This compressor motor causes a-c line voltage fluctuations which are quite rapid and correspond to the cycle of the compressor. Possibly some sort of transients are set up in the eliminator that cause such behavior of the regulator. In record No. 4, however, these variations are not present.

No. 5 is the record of line voltage taken when the eliminator was supplied through the new regulating transformer mentioned above.

The comparison between records Nos. 4 and 5 will readily show that practically no improvement is made by regulating the a-c line voltage through the regulating transformer. As observed from the records, the variation of line voltage from the mean is of the order of 1/4 volt with the new regulator as compared to about 1/2 volt with the older equipment. One advantage of the new regulator, however, is that the variations are not as fast and are quite independent of the load. Thus long periods of constancy are obtained. We have not yet definitely determined why the slow variations seen in the records are more pronounced than those observed when the data for the curves of Figs. 3 and 4 were taken. They seem to be caused either by longitudinal drift of the generator armature in its bearings, or by the application of a large load on one of the 55-volt branches of the line, which brings into operation only one half of the series field.

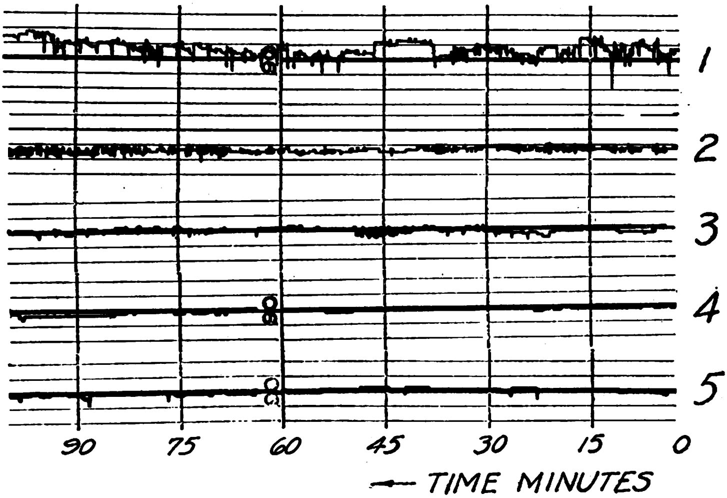

Fig. 5

Simultaneous records of load and voltage were not taken, since only one recorder was at hand. The curves of Figs. 3 and 4, however, clearly show that the new regulator maintains the voltage constant irrespective of the load.

It may be well to mention here that for the sake of experiment an extra stage of amplification was added ahead of the power tube. This introduced hunting and caused the line voltage to oscillate about a mean value at a definite period.

POSSIBLE MODIFICATIONS

This regulator can easily be adapted to line drop compensation and stabilizer installations in which both lighting and elevator loads are present.

A slight modification is necessary in order to apply it to a-c units. By means of a suitable transformer (viz. bell-ringer) the line voltage may be stepped down and rectified by an extra tube placed ahead of the amplifier. The two may then be coupled through a suitable resistance in a somewhat similar manner to that in which the amplifier is coupled to the power tube (Fig. 1). It is obvious that in this case a separate source of power for the B-eliminator is unnecessary. The regulation may therefore be slightly improved, since it is independent of voltage variations of a separate a-c line.

ACKNOWLEDGMENT

The authors are indebted to the Heckscher Research Council of Cornell University, who partially supported this work (grant No. 155). They also wish to thank Prof. Frederick Bedell and Mr. J. G. Kuhn, whose valuable suggestions and interest greatly aided this work.

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Proceedings of the Institute of Radio Engineers
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BOOK REVIEW

Vibration Problems in Engineering. By S. Timoshenko. Published by D. Van Nostrand, 340 pp.

In this book the fundamentals of the theory of vibrations are developed, and their application to the solution of technical problems is illustrated by various examples.

The contents of the book are as follows: the first chapter is devoted to the discussion of harmonic vibrations of systems with one degree of freedom. The general theory of free and forced vibration is discussed, and the application of this theory to balancing machines and vibration recording instruments is shown.

Chapter two contains the theory of the non-harmonic vibration of systems with one degree of freedom. The approximate methods for investigating the free and forced vibrations of such systems are discussed.

In chapter three, systems with several degrees of freedom are considered. The general theory of vibration of such systems is developed and its application in the solution of such engineering problems as: the vibration of vehicles, the torsional vibration of shafts, whirling speeds of shafts on several supports, and vibration absorbers.

Chapter four contains the theory of vibration of elastic bodies. The problems considered are: the longitudinal, torsional, and lateral vibrations of prismatical bars of variable cross section; the theory of vibration of circular rings, membranes, plates, and turbine disks.

Brief descriptions of the most important vibration recording instruments which are of use in the experimental investigation of vibration are given in the appendix.

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Proceedings of the Institute of Radio Engineers
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MONTHLY LIST OF REFERENCES TO CURRENT RADIO LITERATURE

THIS is a monthly list of references prepared by the Bureau of Standards and is intended to cover the more important papers of interest to professional radio engineers which have recently appeared in periodicals, books, etc. The number at the left of each reference classifies the reference by subject in accordance with the scheme presented in "A Decimal Classification of Radio Subjects—An Extension of the Dewey System," Bureau of Standards Circular No. 138, a copy of which may be obtained for 10 cents from the Superintendent of Documents, Government Printing Office, Washington, D.C. The articles listed below are not obtainable from the Government. The various periodicals can be secured from their publishers and can be consulted at large public libraries.

R100. Radio Principles

R007.1 Dellinger, J. H. Engineering aspects of the work of the Federal Radio Commission. Proc. I. R. E., **17**, pp. 1326–1333; August, 1929.

(The engineering problems of the Federal Radio Commission are presented.)

R007.1 Whittemore, L. E. Some principles of broadcast frequency allocation. Proc. I. R. E., **17**, pp. 1343–1353; August, 1929.

(The assignment of available broadcast channels to stations is considered with a view of giving the best possible service to the country as a whole.)

R007.2 Batcheller, A. An outline of the Radio Inspection Service. Proc. I. R. E., **17**, pp. 1365–1376; August, 1929.

(A history of the Radio Inspection Service of the Department of Commerce is given together with an outline of the scope and general nature of its work.)

R113 Krüger, K., and Plendl, H. The propagation of low power short waves in the 1000–kilometer range. Proc. I. R. E., **17**, pp. 1296–1312; August, 1929.

(A description is given of experiments carried out between two ground stations and between an airplane and a ground station to determine the possibility of obtaining reliable short-wave (25 to 55 meters) communication over a distance of 500 km or more with relatively low power (2 watts, continuous wave). An airplane installation operating into a dipole antenna (fixed) is described, and results of a large number of observations at ground receiving stations of the signals received from the airplane in flight are presented.)

R113 Breisig, F. Ueber die Ausbreitung elektrischer Wellen um eine leitende Kugel. (The propagation of electric waves around a conducting sphere). *Elektrische-Nachrichten Technik*, **6**, pp. 268–271; July, 1929.

(The derivation of a formula for finding the field intensity of an electromagnetic radiation at any point on the surface of a conducting sphere is given.)

R113 Eckersley, T. L. An investigation of short waves. *Jnl.* I. E. E., (London), **67**, pp. 992–1032; August, 1929.

(An account is given of investigation made during the past two years of the following wave transmission phenomena: (1) scattering; (2) multiple signals; and (3) signal mutilation of short-wave commercial stations are presented with a discussion of the problems of short-wave transmissions and such cognate subjects as the Heaviside layer, fading, and polarization effects. Certain revisions of results presented in previous papers are offered.)

R113.4 Schelleng, J. C. Further note on the ionization in the upper atmosphere. Proc. I. R. E., **17**, pp. 1313–15; August, 1929.

(Supplement to "Note on the determination of the ionization in the upper atmosphere" which appeared in Proc. I. R. E., **16**, p. 1471; Nov., 1928.)

R113.6 Störmer, C. Kurzwellenechos, die mehrere Sekunden nach dem Hauptsignal eintreffen, und wie Sie sich aus der Theorie des Polarlichtes erklären lassen. (Short-wave echoes that occur several seconds after the main signal and their explanation in the light of the theory of the aurora borealis). *Die Naturwissenschaften,* **17**, pp. 643–51; Aug. 16, 1929.

(Theoretical discussion.)

R125.6 Palmer, L. S. and Honeyball, L. L. K. The action of a reflecting antenna. *Jnl.* I. E. E. (London), **67**, pp. 1045–1051; August, 1929.

(As a study of the action of reflecting antennas the currents produced in two tuned vertical antennas, both acting as receivers in a radiation field, are discussed theoretically. Case 1 treats antennas less than 1 wavelength apart and in line with the transmitter. Case 2 treats antennas less than 1 wavelength apart and at right angles to the transmitter. The conclusions experimentally verified are in the form of an equation, the solutions of which give the critical values of the ratio of the separation distance of the antennas to the wavelength for maximum and minimum antenna currents.)

R125.6 Moser, W. Versuche über Richtantennen bei Kurzen Wellen. (Experiments with short-wave directional antennas). *Zeits. f. Hochfrequenztechnik,* **34**, pp. 19–26; July, 1929.

(A description of experiments and results achieved with several types of short-wave directive antennas.)

R133 Knipping, P. Ueber Barkhausen-Kurz Wellen. (Discussion of Barkhausen-Kurz waves). *Zeits. f. Hochfrequenztechnik,* **34**, pp. 1–12; July, 1929.

(An attempt is made to explain the production of Barkhausen-Kurz oscillations from theoretical considerations of atomic and electronic phenomena.)

R134 Harris, S. An empirical equation for determining the d^2i_g/de_g^2 detectors. Proc. I. R. E., **17**, pp. 1322–25; August, 1929.

(An empirical equation for the grid-current, grid-voltage curve of the C327 electron tube is developed. By obtaining the derivatives of this equation rather than determining them for the i_g-e curve by graphical means the inaccuracy of the graphical method is avoided.)

R134 Warren, A. G. Reduction of distortion in anode rectification. *Experimental Wireless and Wireless Engr.* (London), **6**, pp. 423–37; August, 1929.

(The general principles of plate rectification are reviewed. An analysis is presented to show that distortion due to undesired harmonics can be reduced by plate rectification at high grid voltages.)

R134.6 Hollmann, H. E. Frequenzrückkopplung. (Frequency regeneration). *Elektrische-Nachrichten Technik,* **6**, pp. 253–264; July, 1929.

(Ordinarily regeneration refers to amplitude and not to frequency. However, it is possible to either increase or decrease the frequency of a tube circuit by "frequency regeneration" as described in this article.)

R170 Hogan, J. V. L. A study of heterodyne interference. Proc. I. R. E., **17**, pp. 1354–64; August, 1929.

(The problem of cross talk and heterodyne interference due to the simultaneous use by several broadcasting stations of the same frequency channel or adjacent channels is studied. Strict frequency control and judicious assignments of power and frequency to stations with respect ot their geographical locations are remedies suggested.)

R170 Felix, E. H. The radio engineer's responsibility in coping with man-made interference. Proc. I. R. E., **17**, pp. 1385–89; August, 1929.

(The importance of cooperation of radio interests with power companies and manufacturers of electrical equipment in the elimination of man-made interference is stressed.)

R170 Hooven, M. D., Jr. Radio coordination. Proc. I. R. E., **17**, pp. 1390–94; August, 1929.

(Aspects of the radio interference problem are discussed and the necessity for complete coordination of the numerous interests involved if improvement is to be realized is pointed out.)

R170 RMA better radio reception manual: Home-made static and how to avoid it. Pamphlet by the Radio Mfgrs. Association, Inc., 1929. Price 25 cents.

(A general classification of radio interference is offered and suggestions are given for determining the source of controllable interference. Filter systems and their application to the elimination of certain noises are briefly indicated.)

R200. RADIO MEASUREMENTS AND STANDARDIZATION

R210 Braillard, R. and Devoire, E. The measurement of wavelengths of broadcasting stations. *Experimental Wireless and Wireless Engrs.* (London), **6**, pp. 412–21; August, 1929.

(An account of the work of the Brussels checking station of the U. R. S. I. is presented. The circuit, the operation and the accuracy of the apparatus used in checking the frequency of European broadcasting stations are described.)

R210 Harries, J. H. O. An electromagnetic monochord for the measurement of audio frequencies. PROC. I. R. E., **17**, pp. 1316–21; August, 1929.

(Harmonic monochord theory is outlined and applied in the construction of a secondary audio-frequency standard with the range 32 to 5000 cycles.)

R270 Edwards, S. W. and Brown, J. E. The problem centering about the measurement of field intensity. PROC. I. R. E., **17**, p. 1377–84; Aug., 1929.

(Surveys of service areas of broadcasting stations made by the Radio Division of the Department of Commerce are described. Field intensity contour maps are explained.)

R270 Hollingworth, J. A portable radio intensity measuring apparatus for high frequencies. *Jnl.* I. E. E. (London), **67**, pp. 1033–1044; August, 1929.

(Circuit for apparatus to measure radio field intensities in the range 12000 to 4547 kc (25 to 66 meters) is described. It includes a local oscillator, an attenuator in the form of a resistance-voltage divider, and a heterodyne receiving set and a rectifying unit for measuring the output. An abstract of this paper appeared in *Exp. Wireless and Wireless Engr.* (London) for June, 1929.)

R300. RADIO APPARATUS AND EQUIPMENT

R320 Englund, C. R. and Crawford, A. B. The mutual impedance between adjacent antennas. PROC. I. R. E., **17**, pp. 1277–95; August, 1929.

(Using the methods for treating concealed networks the simple theory for the computation of reflecting antennas is sketched. By experiments based on this theory curves are obtained showing the relation of the phase angle between the currents in the transmitting antenna and in the reflecting antenna to the distance in wavelengths between them for the range $\lambda/3$ to infinity where $\lambda = 4.34$ meters.

R330 Jouaust, R. L'etat actuel de la technique des lampes a plusieurs electrodes. (The present state of the art of multi-electrode vacuum tubes). *L'Onde Electrique*, **8**, pp. 227–261; June, 1929.

(The results and formulas acquired through progress in vacuum-tube art are assembled and reviewed. Interelectrode capacity, secondary emission, the phenomenon of the virtual cathode are treated. The multi-grid and the heater type tubes are included in the review.)

R331 Decaux, B. Les caracteristiques des lampes de reception modernes et leur choix rationnel. (Characteristics of modern receiving tubes and their rational choice). *L'Onde Electrique*, **8**, pp. 262–280; June, 1929.

(A classification of various types of tubes according to their amplification factors and their plate resistances is suggested as a useful guide to users in the choice of a tube for a specific use. The characteristics of multi-grid tubes and the interesting results they offer are noted.)

R343 von Handel, P; Krüger, K; Plendl, H. Quarzsteuerung von Kurzwellen Empfängern. (Crystal control for short-wave receiving sets). *Zeits. f. Hochfrequenztechnik*, **34**, pp. 12–18; July, 1929.

(Describes several methods of using crystal control in short-wave receiving sets, the most successful being the crystal controlling of the oscillations of a regenerative detector tube which forms an intermediate frequency with the incoming signal. The intermediate frequency is then amplified and detected.)

R360 Hull, L. M. Some characteristics of modern radio receivers and their relation to broadcast regulation. PROC. I. R. E., **17**, pp. 1334–41; August, 1929.

(A report is made of comparative tests on twenty 1927–1928 receiving sets and on twenty-four 1928-1929 receiving sets. The sets are compared with respect to (1) their discrimination between channels; (2) their discrimination within a channel; (3) their uniformity of reception in all channels; (4) their uniformity of reception within a channel: and (5) their range.)

R376.3 Meyer, E. and Just, P. Frequenzkurven von elektrischen Tonabnehmern und mechanischen Grammophonen. (Frequency curves of electrical reproducers and mechanical gramaphones). *Elektrische-Nachrichten Technik*, **6**, pp. 264–68; July, 1929.

(Curves are presented.)

R500. APPLICATIONS OF RADIO

R550 Marriott, R. H. United States radio broadcasting development. PROC. I. R. E. **17**, pp. 1395–1439; August, 1929.

(A detailed account of the development of radio broadcasting in the United States from 1907 to 1928 inclusive; followed by a description of the route of a modern broadcast from the studio to the listener. Possible future developments are discussed.)

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Little, D. G.: Born 1893 at Kalamazoo, Michigan. Attended Kalamazoo College; received B.S. degree in engineering, University of Michigan, 1917. Second lieutenant, U. S. Army, 1917–1919; assigned radio development work at U. S. Bureau of Standards and signal corps laboratory at Camp Alfred Vail. With Westinghouse Electric and Manufacturing Co., 1919 to date. Chief engineer, Chicopee Falls Works, 1927 to date. Associate member, Institute of Radio Engineers, 1920; Member, 1926.

Metcalf, G. F.: Received B.S. degree, Purdue University, 1928. Research laboratory, General Electric Co., 1928 to date.

Nelson, Edward L.: See PROCEEDINGS for August, 1929.

Reich, Herbert J.: Born October 25, 1900 at New Dorp, N. Y. Received M.E. degree, Cornell University, 1924. Instructor, machine design and physics, Cornell University, 1924–1929; Ph.D. degree, 1928. Assistant professor of electrical engineering, University of Illinois, beginning September 1, 1929. Associate member, Institute of Radio Engineers, 1926; member of several other societies.

Verman, Lal Chand: Born October 2, 1902 at Amritsar, India. Received B.S. degree in electrical engineering, University of Michigan, 1927; M.S. degree, Cornell University, 1928. Research assistant in engineering research department and Donnovan scholar, University of Michigan, 1925–1927. Charles A. Coffin fellow, Cornell University, 1927–1929. Summers at Detroit Edison Co. At present Heckscher Research assistant and candidate for Ph.D. at Cornell University. Associate member, American Institute of Electrical Engineers. Associate member, Institute of Radio Engineers, 1928.

Weinberger, Julius: See PROCEEDINGS for September, 1929.

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VOLUME 17 DECEMBER, 1929 NUMBER 12

PROCEEDINGS
of
The Institute of Radio Engineers

General Information and Subscription Rates on Page 2090

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PROCEEDINGS OF
The Institute of Radio Engineers

Volume 17 December, 1929 Number 12

Board of Editors, 1929
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RALPH BATCHER L. E. WHITTEMORE
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The Institute of Radio Engineers

GENERAL INFORMATION

The PROCEEDINGS of the Institute is published monthly and contains papers and discussions thereon submitted for publication or for presentation before meetings of the Institute or its Sections. Payment of the annual dues by a member entitles him to one copy of each number of the PROCEEDINGS issued during the period of his membership.

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Year Books for 1926, 1927, and 1928, containing general information, the Constitution and By-Laws, catalog of membership etc., are priced at seventy-five cents per copy per year.

Contributors to the PROCEEDINGS are referred to the following page for suggestions as to approved methods of preparing manuscripts for publication in the PROCEEDINGS.

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Changes of address to affect a particular issue must be received at the Institute office not later than the 15th of the month preceding date of issue. That is, a change in mailing address to be effective with the October issue of the PROCEEDINGS must be received by not later than September 15th. Members of the Institute are requested to advise the Secretary of any change in their business connection or title irrespective of change in their mailing address, for the purpose of keeping the Year Book membership catalog up to date.

The right to reprint limited portions or abstracts of the papers, discussions, or editorial notes in the PROCEEDINGS is granted on the express condition that specific reference shall be made to the source of such material. Diagrams and photographs published in the PROCEEDINGS may not be reproduced without making special arrangements with the Institute through the Secretary.

It is understood that the statements and opinions given in the PROCEEDINGS are views of the individual members to whom they are credited, and are not binding on the membership of the Institute as a whole.

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SUGGESTIONS FOR CONTRIBUTORS TO THE PROCEEDINGS

Preparation of Paper

Form—Manuscripts may be submitted by member and non-member contributors from any country. To be acceptable manuscripts should be in final form for publication and accompanied by a summary of from 100 to 300 words. Papers should be typed double space with consecutive numbering of pages. Footnote references should be consecutively numbered and should appear at the foot of their respective pages. Each reference should contain author's name, title of article, name of journal, volume, page, month, and year. Generally, the sequence of presentation should be as follows: statement of problem; review of the subject in which the scope, object, and conclusions of previous investigations in the same field are covered; main body describing the apparatus, experiments, theoretical work, and results used in reaching the conclusions and their relation to present theory and practice; bibliography. The above pertains to the usual type of paper. To whatever type a contribution may belong, a close conformity to the spirit of these suggestions is recommended.

Illustrations—Use only jet black ink on white paper or tracing cloth. Cross-section paper used for graphs should not have more than four lines per inch. If finer ruled paper is used, the major division lines should be drawn in with black ink, omitting the finer divisions. In the latter case, only blue-lined paper can be accepted. Photographs must be very distinct, and must be printed on glossy white paper. Blueprinted illustrations of any kind cannot be used. All lettering should be $^{3}/_{16}$ in. high for an 8 x 10 in. figure. Legends for figures should be tabulated on a separate sheet, not lettered on the illustrations.

Mathematics—Fractions should be indicated by a slanting line. Use standard symbols. Decimals not preceded by whole numbers should be preceded by zero, as 0.016. Equations may be written in ink with subscript numbers, radicals, etc., in the desired proportion.

Abbreviations—Write a.c. and d.c., (a-c and d-c as adjectives), kc, μf, $\mu\mu$f, e.m.f., mh, μh, henries abscissas, antennas. Refer to figures as Fig. 1, Figs. 3 and 4, and to equations as (5). Number equations on the right in parentheses.

Summary—The summary should contain a statement of major conclusions reached, since summaries in many cases constitute the only source of information used in compiling scientific reference indexes. Abstracts printed in other journals, especially foreign, in most cases consist of summaries from published papers. The summary should explain as adequately as possible the major conclusions to a non-specialist in the subject. The summary should contain from 100 to 300 words, depending on the length of the paper.

Publication of Paper

Disposition—All manuscripts should be addressed to the Institute of Radio Engineers, 33 West 39th Street, New York City. They will be examined by the Committee on Meetings and Papers and by the Editor. Authors are advised as promptly as possible of the action taken, usually within one month.

Proofs—Galley proof is sent to the author. Only necessary corrections in typography should be made. *No new material is to be added.* Corrected proofs should be returned *promptly* to the Institute of Radio Engineers, 33 West 39th Street, New York City.

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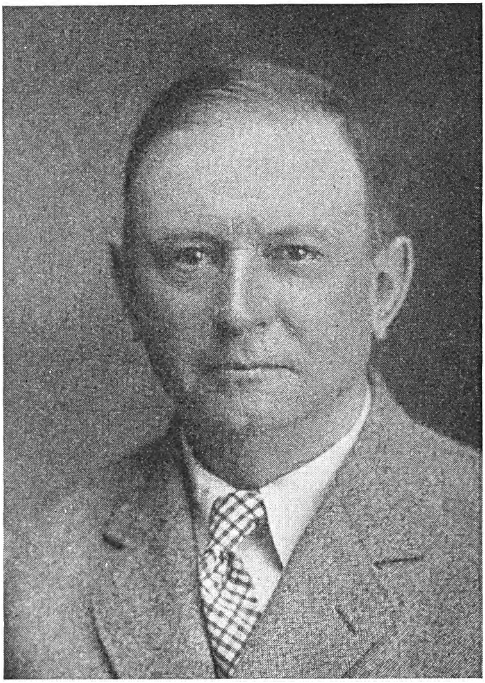

George Washington Pierce
President of the Institute, 1918 and 1919

George Washington Pierce was born in Webberville, Texas, January 11, 1872. He received the B.Sc. degree, University of Texas, 1893; M.A. degree, 1894; Harvard, 1898–1900, A.M. degree, 1899, Ph.D. degree, 1900; Leipzig, 1900–01. Asst. professor of physics, 1907–17; professor, 1917–21; Rumford professor, 1921 to date, Harvard University. Director of the Cruft High Tension Electrical Laboratory, 1914 to date; chairman of the Division of Physical Sciences, Harvard, 1927 to date. Fellow of the American Academy of Arts and Sciences; member of the National Academy of Sciences, American Physical Society, American Institute of Electrical Engineers, Deutsche Physikalische Gesellschaft, American Acoustical Society, Inventors' Guild; Fellow and twice president of the Institute of Radio Engineers. Author of "The Principles of Wireless Telegraphy," 1910; "Electric Oscillations and Electric Waves," 1920; and of papers on crystal rectifiers, sound, electrical resonance, electric waves, transmission lines, theory of the antenna, piezo-electric crystal control of frequency, and applications of magnetostriction to frequency control and to the production and detection of sound. Inventor of devices for submarine signalling, vacuum-tube detectors, frequency control apparatus, etc.

INSTITUTE NEWS AND RADIO NOTES

November Meeting of Board of Direction

The regular monthly meeting of the Board of Direction of the Institute was held on November 7, 1929 in the offices of the Institute.

Those present were: W. G. Cady, acting chairman; Melville Eastham, treasurer; John M.Clayton, secretary; Arthur Batcheller, R. A. Heising, L. M. Hull, R. H. Manson, and R. H. Marriott.

The following were transferred or elected to the higher grades of membership in the Institute:

Transferred to the Fellow grade: R. H. Langley, W. D. Terrell, and Balth. van der Pol. Transferred to the Member grade: C. S. Agate, G. L. Beers, H. A. Brown, I. S. Coggeshall, H. I. Danziger, H. B. Deal, W. F. Diehl, Malcolm Ferris, H. C. Forbes, J. K. Hilliard, A. V. Loughren, J. R. Nelson, E. B. Ferrell, and O. W. Towner. Elected to the Member grade: Arthur S. Fish, J. Plebanski, T. C. Rives, and D. M. Sokolcow.

One hundred and thirteen Associate members and six Junior members were elected.

K. W. Wagner, president of the Electrotechnical Association, with headquarters at Berlin, invited the Institute to appoint a representative to the fiftieth anniversary celebration of the Association to be held in Berlin in January, 1930. The Board accepted the invitation, nominating Vice-President Alexander Meissner as the Institute's representative.

Melville Eastham has been appointed a member of the Radio Broadcasting Committee of the American Engineering Council.

A Request from the Committee on Admissions

The Institute of Radio Engineers has attained a high standing among engineering societies through the quality of its publications and the calibre of its membership. The Committee on Admissions is charged with the important responsibility of maintaining proper standards of membership. The Committee feels it to be important to every member of the Institute that the standards of membership, set by the Constitution, be maintained.

The Committee on Admissions occasionally experiences difficulty in passing upon applications for the higher grades of membership because of great reluctance on the part of many references to state frankly and fully their knowledge and opinions concerning the applicant. There seems to be a disinclination to say anything about the applicant unless he can be recommended highly. It is in such uncertain cases that the Committee particularly needs the assistance of every reference. The Committee cannot take the failure to return the reference blank as non-indorsement since there are many cases where such a course would do great injustice to the applicant. This condition does not hold among our British members. A healthy critical attitude prevails that has been of great help to the Committee. This desirable attitude is prevalent in England and has been instrumental in maintaining the high standards of their engineering societies. The development of such an attitude is absolutely necessary if the Institute wishes to maintain or improve its standards of membership. The Committee earnestly requests all references to be sure to reply and, for the good of the Institute, to express themselves frankly and honestly regardless of personal or business relations. All replies from references are kept in the strictest confidence by the Committee.

January, 1930, New York Meeting

Since the normal date of the New York meeting in January next falls on the first of the month, this meeting will be changed to Wednesday, January 8th.

1930 Membership Dues

It is planned that bills for members' dues for 1930 will be in the mails shortly before January 1st. Prompt payment of these bills will insure the regular receipt of the PROCEEDINGS and other Institute publications, and, due to the small office staff and limited bookkeeping facilities, will be of great assistance to the Institute office.

Change in Business Title

Members are advised that any change in their business title from that shown in the 1929 Year Book should be communicated to the Secretary of the Institute immediately if he has not been advised thereof previously. The catalog of membership in the 1930 Year Book comprises the names and addresses of the paid membership as of

December 31, 1929. Changes in mailing address, business address, or business title will not appear in the 1930 book unless received at the Institute office by December 31st. A form for notification of change in address is published monthly in the advertising section of the PROCEEDINGS. It will be found on page xxxv of this issue.

Radio Standardization

While the members of the Institute know that a considerable amount of standardization work is in progress, they may not have a very extensive knowledge of the important part the Institute is playing in this field.

There are, in general, four major bodies keenly interested in standardization in the radio field. These are the American Standards Association, the Institute of Radio Engineers, the Radio Division of the National Electrical Manufacturers' Association, and the Radio Manufacturers' Association.

In 1918, five of the major engineering societies decided it would be extremely desirable to have an organization whose chief duties it would be to encourage and coordinate standardization in the engineering field, to determine upon suitable methods of procedure which would entail the minimum loss of time and effort commensurate with a high standard of accuracy and suitability of the standards formulated, and which would maintain sufficient supervision of the process of developing and promulgating standards to insure that those which received its approval were truly representative of all those in the industry having a vital interest in the subject. With these thoughts in mind, the American Engineering Standards Committee was organized.

As time passed, other organizations became affiliated with the A. E. S. C. and the scope of its activities naturally broadened until, in 1928, it was considered advisable to reorganize the committee along much broader lines which would also allow for future expansion. In this reorganization, the name was changed to the American Standards Association.

At the present time the A. S. A. has forty-one member-bodies which comprise technical societies, trade associations, and governmental bodies, all of which are keenly interested in the problems confronting a large number of industries.

The broad scope of the work may be readily demonstrated by going through the list of projects for standardization under the A. S. A.

procedure. These concern such items as Portland cement, tests for the toughness of rock, safety codes for ladders and for window washing, dimensional standards for screws and pipe threads, gears, refrigerators, radio, power-line insulators, mercury-arc rectifiers, storage batteries, street traffic signs, liquid soap, fire hose, bed sheeting, blankets, mine timbering, wood poles, preferred numbers, and many others.

The A. S. A. radio standardization program is sponsored by the Institute of Radio Engineers and the American Institute of Electrical Engineers, jointly, through the Sectional Committee on Radio which has for the scope of its work, "nomenclature, methods of testing and of rating, specifications of apparatus and equipment, and dimensions to secure interchangeability where this may be found desirable."

One of the most important duties of the A. S. A. is to scan carefully the personnel of all its sectional committees to insure that there is a suitable balance of power among the producers, distributors, consumers, and general interests. In compliance with this requirement, it will be found that practically every important organization in the radio industry in this country is represented on the Sectional Committee on Radio.

In granting its approval of a standard, the A. S. A. does not concern itself with the content and merits of the standard, but guarantees that it is the consensus of opinion of all who are or should be vitally interested in the subject.

As a result of unavoidable conditions existing in the organization in the past, the work of the Sectional Committee on Radio was considerably hampered, and culminated in the acceptance by the A. S. A. of but one "American Tentative Standard." This covers the "Specifications for Vacuum-Tube Bases" and has already been published in the PROCEEDINGS. However, the Sectional Committee on Radio has recently been completely reorganized, radically revising its plan of work, and it is expected that the results of its operation during 1929 and 1930 will be materially greater.

In addition to its sponsorship of the Sectional Committee on Radio, the Institute has its own Committee on Standardization which has done effective work for a number of years. Its efforts to date have been confined to matters of nomenclature, definitions, symbols, and methods of rating and of testing equipment and apparatus. In this way, it has expended its energies on such matters as are of greatest importance to radio engineers.

In the manufacturing field, the Radio Division of the National Electrical Manufacturers' Association and the Radio Manufacturers' Association have each sponsored a program of standardization. They have made no attempt to formulate standards in that portion of the field already covered by the Institute, but have taken these standards bodily from the Institute's reports and given them their own approval as well. These standards, therefore, have the united approval of the Institute and the two major manufacturers' associations. It is anticipated that a goodly portion of these will be accepted by the Sectional Committee on Radio during the ensuing year. They will then be presented to the A. S. A. for approval.

The programs of the N. E. M. A. and R. M. A. have been confined to matters of concern in the manufacture of equipment, particularly such items as are of importance as regards the size and fit of component parts to secure interchangeability, markings for identification, which include such things as color codes for wiring, and terminal marking for polarity, and many matters of similar nature.

It will thus be seen that the field has been divided into two parts, the Institute providing for the more fundamental engineering aspects as regards nomenclature, methods of test, etc., and the manufacturers' associations assuming the obligation of providing standards covering interchangeability of parts and such matters of a more distinctly manufacturing nature.

Because of the nature of the products sold to the general public, it is not customary for identifying labels to be appended to items complying with formulated standards. Neither is there any coercive force applied to compel compliance with these standards; general recognition of their suitability and excellence is relied upon to encourage their acceptance by the industry. This procedure has met with good success and is but a continuation of the methods employed in generation of the standards.

These standards have made it possible for the manufacturer and those from whom he purchases materials and those to whom he, in turn, sells, to speak the same language as regards nomenclature, ratings, and methods of test. In this way more accurate computation of costs, not only of component parts but of the finished receiver as well, may be made.

By standardizing on the dimensions of component parts, the manufacturer is enabled to dispose of his stock to a large number of purchasers who can all use the same item but who, under less favorable conditions to the manufacturer, would demand their own particular sizes, shapes, etc., thus increasing the manufacturing cost per unit. It also allows the manufacture of materials and parts during the less active portions of the year under conditions which will not make it impossible to find a reasonable market for them when the busier months arrive and time and materials are at a premium.

The advantages and benefits to the industry from standardization are manifold, and practically all result in a general saving to the ultimate consumer in the cost and ease of servicing and maintaining his receiving set and in lowered cost of construction, installation, and maintenance of broadcast stations, which, in turn, make his receiver so desirable. This lowered cost and improved service tend to encourage those who have not already purchased receivers to do so, making a larger total volume of business and a further reduction in the cost per unit.

Standardization, therefore, helps not only the ultimate consumer but also the manufacturer, distributor, and all others in the industry. By its means, industries are encouraged to grow—and continue to grow, and the savings made become available for new research and for improvements in the methods and processes.

Meeting of the A. A. A. S. in December

Under the chairmanship of Professor J. C. Jensen of Nebraska Wesleyan University a committee of Institute members has been appointed to arrange a program for Institute members at the meeting of the American Association for the Advancement of Science which will be held in Des Moines, Iowa, December 27 to 31, 1929. The committee membership follows: J. C. Jensen, chairman; S. T. Hutchinson, Charles A. Culver, H. M. Carrouthers, and John H. Miller.

It is expected that several papers on radio subjects will be presented at the Institute meeting. Members desiring further information should communicate either with the Institute office or Professor Jensen.

Institute Meetings

Tour to Sections by K. S. Weaver

During the month of October, K. S. Weaver of the Westinghouse Lamp Company of Bloomfield, N. J., visited five sections of the Institute to deliver the paper, "Production Testing of Vacuum Tubes," by W. S. Jones and himself.

The meetings in each section were arranged as follows:

Chicago Section.—On October 18th, in the Electric Club, Chicago. G. W. Wilcox presided. Seventy members of the section were present. The paper was discussed by Messrs. Wilcox, Oxner, Miller Harper, and others.

Cincinnati Section.—On October 17th in the engineering building, University of Cincinnati. R. H. Langley, chairman of the section, presided. Seventy-four members and guests attended. Messrs. Kigour, Felix, Israel, and others participated in the discussion of the paper.

Connecticut Valley Section.—On October 14th in the auditorium of the Hartford Electric Light Company. Thirty-seven members were present.

Detroit Section.—On October 16th in the conference room of the Detroit News building. A. B. Buchanan, chairman of the section, presided. Seventy-five members and guests were present. The paper was largely discussed.

Toronto Section.—On October 15th in the Westinghouse Auditorium at Hamilton, Ontario. V. G. Smith, chairman of the section, presided. One hundred and twenty-five members and guests were present. The paper was discussed by Messrs. Patience, Mitchell, Thompson, and others.

This paper, which will be published in a forthcoming issue of the PROCEEDINGS, is summarized as follows:

> "The factory testing of vacuum tubes is becoming more specialized every year. In the early days of broadcasting a few tubes served many purposes and actual performance in its relation to tube characteristics was not as widely studied and understood as it is today.
>
> "Increasing knowledge of the 'how' and 'why' of tube and set operation has led to the development of many more or less special tubes designed for a specific function in the set. The increasing use of accurate set measuring equipment has enabled designers to work out the optimum circuit arrangements for the use of these specially designed tubes in the various stages.
>
> "As a result of this growth of the industry the testing of vacuum tubes has required more and more attention to set operation, a constant elaboration of test methods and equipment being necessary in order to exercise the necessary control over tube characteristics."

CHICAGO SECTION

The regular meeting of the Chicago section for the month of November was held on the fourth in the Engineering Hall, Engineering Building, Chicago. John H. Miller, chairman, Telegraph, Telephone and Radio Section, Western Society of Engineers, and secretary of the Chicago section, presided.

M. Merwin Eells, manager of communications of National Air Transport, presented a paper, "Radiobeacon Systems for Aircraft." The paper described in some detail the work of National Air Transport with aural beacons.

Preceding the paper's presentation a reel of motion pictures depicting the radiobeacon system and airplane radio systems described in the paper was shown.

This was a joint meeting with the Telephone, Telegraph and Radio Section of the Western Society of Engineers. One hundred and fifty-seven members and guests were present.

CLEVELAND SECTION

On November first the Cleveland section held a meeting in Case School of Applied Science. Bruce W. David, chairman of the section, presided.

Herbert A. Erf presented an illustrated talk on "Acoustics and Radio Engineering." After giving an historical setting to the study of acoustics the speaker discussed such topics as interference, resonance, echo, and reverberation. In the general discussion which followed Mr. Erf's paper many interesting points were brought out in connection with acoustical design of broadcast studios.

In the business meeting preceding the presentation of the paper the following nominating committee was appointed to make nominations for 1930 section officers: J. R. Martin, Ralph Farnham, and G. B. Schneeberger.

LOS ANGELES SECTION

On October 21st a meeting of the Los Angeles section was held at Fox Hills Studio, Beverly Hills. A dinner was served at the Cafe

Institute of
Radio Engineers
Fox Hills Studio
Oct. 21, 1929
Moss
Photo
L.A.

DINNER OF LOS ANGELES SECTION AT FOX HILLS STUDIO, OCTOBER 21, 1929 de Paris within the studio grounds. During the dinner hour, music was furnished by a portable equipment. H. Keith Weeks, executive manager of Fox Studios, introduced the speaker of the evening, K. E. Morgan, who gave a very interesting talk on the subject, "Electrical Engineering of Sound Picture Systems." Mr. Morgan is associated with the Electrical Research Products Corporation.

The talk was illustrated by means of a complete portable movietone projector showing a set of stereopticon slides and three especially prepared reels of motion pictures with sound effects.

Following the talk those present visited one of the new sound stages where the various equipment in the stage, control room, and recording room was explained in detail and shown in operation.

Four hundred and eighty members and guests attended this meeting. A photograph of the dinner gathering can be obtained from the secretary of the section, W. W. Lindsay, at a nominal sum.

NEW ORLEANS SECTION

A meeting of the New Orleans section was held on October 10 at Loyola University, New Orleans. J. N. Dutreil presided. Two motion pictures, "Man-made Miracles" and "The Nation's Market Place," were shown. Following the meeting members inspected the 5-kw broadcast station operated by the University.

NEW YORK MEETING

The regular monthly New York meeting was held on November 6 in the Engineering Societies Building, 33 West 39th Street. In the absence of President Taylor, W. G. Cady presided.

The first of two papers presented was divided into two parts. The first part was read by I. F. Byrnes, of General Electric Company, Schenectady, and the second part by J. B. Coleman, Westinghouse Electric and Manufacturing Company, Chicopee Falls. Messrs. Coleman and Byrnes were the joint authors of the paper entitled "Short-Wave Communication." The paper is summarized as follows:

> "One of the major parts of the short-wave communication system is the transmitter converting 60-cycle power into a frequency in the order of millions of cycles. Although the first communication work on short waves was accomplished by the radiation of a few watts, reliable and rapid communication requires power of the order of tens of kilowatts. The technique in producing a transmitter of relatively large power to feed an antenna for radiating short waves, has been found to be quite different from that required for previous design and test of radio transmitters. This paper presents the results of intensive work of the last two years. It describes some of the interesting problems which have had to be solved. Not least

> among these were methods of providing a satisfactory artificial load, the determination of keying characteristics and the like. Illustrations showed the concrete form of the equipment giving an interesting comparison between present communication equipment and its immediate predecessor, the Alexanderson alternator."

The second paper, "Some Problems in Short-Wave Telephone Transmission," by J. C. Schelleng of Bell Telephone Laboratories, New York, was presented by Mr. Schelleng. This paper is summarized as follows:

> "Certain phases of short-wave telephony, primarily, though not entirely, from the point of view of the transmitter are described. The field strengths which the transmitting station must provide at the receiver are considered. Typical data are given showing results obtained in transmission from Deal, N. J., to England. This is followed by a discussion of requirements and limitations of the transmitting antenna. The gains which arrays may reasonably be expected to provide are considered. The phenomenon of non-synchronous fading at nearby points is examined as to its bearing on the dimensions and performance of directive arrays. It is concluded that in a broadside antenna a loss of about 1 db is entailed when the length of the structure equals the minimum distance of random fading. Other directional properties of the transmitting medium are also considered. Attention is then directed to the transmitting equipment, particular attention being given to the high power part of it. Requirements, rather than circuit details, are emphasized. These include stability of operation, flexibility and freedom from amplitude distortion, and phase and frequency modulation. The latter two types of distortion, while having much in common, arise in different manners and have certain distinguishing characteristics. The results of tests in which some of these matters were considered quantitatively are given."

It is expected that both of these papers will appear in a forthcoming issue of the PROCEEDINGS.

Four hundred and fifty-five members of the Institute attended the meeting.

PHILADELPHIA SECTION

The Philadelphia section met on October 3 in the Franklin Institute, 15 S. 7th Street.

J. C. Van Horn, chairman of the section, presided. Elmer L. Brown, of the Engineering and Test Laboratory, Radio-Victor Coporation of America, presented a paper, "The Progress of Aircraft Radio." This paper constituted a resumé of aircraft radio engineering from the beginning of the World War to the present date.

The rigid requirements for this type of apparatus, specific problems encountered, special apparatus developed specifically for aircraft needs and the principles of popular types of altimeters and beacons were discussed. The trend of future development was suggested.

Messrs. Kenrick, Snyder, and others participated in the discussion which followed the presentation of the paper.

Sixty-four members of the section and guests attended the meeting.

On November 5th a meeting of the section was held in the Franklin Institute, Philadelphia. J. C. Van Horn, chairman of the section, presided.

Stuart Ballantine, of Boonton Research Corporation, presented a paper on "Recent Developments in RFL Broadcast Receivers." This paper was, in a sense, a continuation of one presented by the present author at a meeting of the Philadelphia section on May 21, 1926, entitled "Recent Developments in Radio Receivers." The present paper included a historical review of RFL receiver developments from 1922 to 1927, particularly the development of monodic (one-way or neutralized) radio-frequency amplifiers, single-tuning control, and complete shielding. Methods of measuring the overall electrical performance of a radio receiver as indicated by the sensitivity, fidelity, and selectivity were described. Then followed a discussion by Mr. Ballantine of more recent research projects of commercial interest including the use of shielded tetrodes in r-f amplifiers, volume control, detection by diode, grid and plate rectification at high signal levels, the design procedure in such systems, preventing of detector overloading, linear detectors, distortion in detection due to high modulation, systems for automatic volume control, and the use of a pentode in the power output stage.

Two hundred and fifteen members attended the meeting.

PITTSBURGH SECTION

The Pittsburgh section met in the Chamber of Commerce Building, Pittsburgh, on October 15th jointly with the Pittsburgh section of the A. I. E. E. J. A. Cadwallader, chairman of the latter section, presided.

John B. Taylor, of the Research Laboratory, General Electric Company, presented a lecture on "Making Sound Visible and Light Audible."

Seven hundred members of the two sections were present.

SAN FRANCISCO SECTION

On October 23rd the San Francisco section met in the Engineers' Club, 206 Sansome Street, San Francisco. Donald K. Lippincott, chairman of the section, presided.

Dr. Lester E. Reukema presented a lecture on "Radio Research and Research Men in Europe," describing his recent visit to research laboratories in England, France, Germany, Austria, Italy, Sicily, North Africa, Switzerland, and Holland. He related an account of various research problems under way in laboratories in these countries.

Fifty-one members attended the meeting.

SEATTLE SECTION

A meeting of the Seattle section was held on October 25th in Philosophy Hall, University of Washington. Austin V. Eastman, chairman of the section, presided.

James Gordon Bennett, radio electrician, Airways Division, Department of Commerce, presented a paper, "Airways Division Organization, Operation, and Service Rendered." The paper pointed out that the Airways Division was organized in 1926 after some fourteen years of consideration and investigation. It was established to aid airplane navigation, as at that time there were no facilities for regulation and information. Some of the duties of the airways division are the testing and inspecting of the new ships, the testing and inspecting of engines, the erection and maintenance of beacon lights and radio stations to send out weather reports and other information of use to airplane pilots.

Fifty-nine members of the section attended the meeting.

WASHINGTON SECTION

On October 10th in the Continental Hotel, North Capitol Street, Washington, a meeting of the Washington section was held. Thomas McL. Davis, vice-chairman of the section, presided.

R. H. Ranger, of the Radio Corporation of America, presented a talk on "Photoradio." His paper is summarized as follows:

> "The development of an accurate system of photoradio successfully used between London, New York, and San Francisco is described and the circuits by which multiplex facsimile transmission and reception are accomplished is described. The talk was illustrated by lantern slides showing the different circuits employed for multiplex facsimile transmission. The method of electrically commutating the several multiplex channels in lieu of mechanical distributor separation employed in radiotelegraph circuits was explained. The fact that no mechanical device is employed in the multiplex transmission and reception system for opening and closing the circuit permits higher speeds and more accurate operation of the circuits to be obtained. The multiplex system involves separate banks of electron tube amplifiers, the number of banks corresponding to the number of channels. These amplifiers are associated with independent transmission circuits and permit these circuits to function in successive order by electrical commuta-

> tion; that is, the tube circuits of the different amplifier banks are interconnected so that when one bank is functioning the grids of the other banks are positive, and as the bank operates in successive order the grids are made successively positive or negative by a tripped system, or released by the circuits in adjacent banks."

The discussions of the paper were made by Messrs. Taylor, Gunn, Mirick, Smith, Brady, Davis, and others.

Seventy-five members and guests attended this meeting.

Committee Work

Committee on Admissions

A meeting of the Committee on Admissions was held on November 6th in the Institute office. Those present were: R. A. Heising, chairman; E. R. Shute, A. F. Van Dyck, and H. F. Dart. The committee considered twenty-two applications for transfer or election to higher grades of membership.

Subcommittee on Receivers and Parts

The Subcommittee on Receivers and Parts of the Institute's Committee on Standardization for 1929 held its first meeting in the Institute office on the morning of November 6th. Members present were: E. T. Dickey, chairman; G. C. Crom, C. E. Brigham, W. A. MacDonald, C. M. Sinnett, V. M. Graham, E. J. T. Moore, I. G. Maloff, and H. P. Westman, secretary.

Subcommittee on Transmitters and Antennas

The Subcommittee on Transmitters and Antennas of the Committee on Standardization held its first meeting in the Institute office on November 7th. Those present were: Haraden Pratt, chairman; R. M. Wilmotte, H. E. Hallborg, J. J. Lamb, and H. P. Westman, secretary.

Board of Editors

A meeting of the Board of Editors of the Proceedings was held the afternoon of November 6th. The following members of the Board were present: W. G. Cady, chairman; Ralph Batcher, and L. E. Whittemore.

The Board decided upon several typographical changes to the Proceedings to be effective with the January, 1930, issue. Many questions relative to policies were discussed.

SECTIONAL COMMITTEE ON RADIO

At a meeting on October 7th, 1929, in the office of the Institute, the reorganization of the Sectional Committee on Radio of the American Standards Association, of which the Institute and the A. I. E. E. are joint sponsors, was undertaken.

Officers of the new Committee were elected as follows: Alfred N. Goldsmith, chairman; C. H. Sharp, vice chairman; and H. P. Westman, secretary.

The membership of the reorganized Sectional Committee and the organizations represented is as follows: Lloyd Espenchied, Bell Telephone System; L. W. Chubb, W. R. G. Baker, and R. H. Manson, National Electrical Manufacturers' Association; Alfred N. Goldsmith, Radio Corporation of America; W. E. Holland, R. H. Manson, and R. H. Langley, Radio Manufacturers' Association; J. J. Lamb, American Radio Relay League; R. H. Langley, National Association of Broadcasters; J. J. Graf, Telephone and Telegraph Section, American Railway Association; T. A. M. Craven, United States Navy Department; F. W. Hoorn, United States War Department; C. H. Sharp, Adam Stein, Jr., and William Wilson, American Institute of Electrical Engineers; L. E. Whittemore, J. V. L. Hogan, and L. M. Hull, Institute of Radio Engineers; R. N. Conwell, National Electric Light Association; R. B. Shepard, National Fire Protection Association and Underwriters' Laboratories; E. W. Ely, United States Department of Commerce; J. H. Dellinger, United States Interdepartment Radio Advisory Committee; and O. H. Caldwell, at large.

As chairmen of the Technical Committees the following were appointed: Haraden Pratt, radio transmitters and parts; W. A. MacDonald, radio receivers and parts; J. C. Warner, vacuum tubes; Irving Wolff, electro-acoustic devices.

Personal Mention

J. E. Anderson, formerly radio engineer of Commercial Radio Service Company of Columbus, Ohio, has recently become laboratory assistant at Day Fan Electric Company of Dayton, Ohio.

Major Arthur L. Harris has been transferred from Royal Signals, War Office, London, to China Command Signal Company at Hong Kong, China.

J. R. Harrison, until recently assistant in physics department of Wesleyan University at Middletown, Conn., is now on the staff of the Department of Physics, University of Pittsburgh.

Clifford E. Himoe has resigned from the operating staff of Station WEAF at Bellmore, L. I., to become associated with DeForest Radio Company of Passaic, N. J., in the engineering laboratories.

Commander C. H. Maddox has been transferred from the *USS Putnam* to the Navy Department, Washington, D. C.

B. F. McNamee, until recently chief engineer of Advance Electric Company of Los Angeles, California, is now employed as chief production engineer with Colin B. Kennedy Radio Corporation at South Bend, Indiana.

H. P. Miller, Jr., has resigned from Federal Telegraph Company, Newark, N. J., where he was employed as radio transmission engineer to become radio engineer in International Communications Laboratories, New York City.

A. F. Murray has joined the engineering department of Jenkins Television Laboratory at Jersey City, N. J. Mr. Murray was formerly associated with Wireless Specialty Apparatus Company of Boston.

Alan N. Ramsay, until recently general sales manager of Precision Electric Manufacturing Corporation of Los Angeles, is now production engineer, projection department, Vitovox Talking Pictures Company at Los Angeles.

Philip A. Richards, for the past two years in the radio department of General Electric Company at Schenectady, has been transferred to the Nela Park office as vacuum-tube engineer.

R. Oliver Rippere, recently student at Brooklyn Polytechnic Institute, is now in the telephone engineering department of Bell Telephone Laboratories.

Arthur W. Steinberger, a former acoustic engineer with Charles Freshman Company, is now employed by Colonial Radio Corporation, Long Island City, as acoustic engineer.

C. M. Jansky, Jr., until recently associate professor, radio engineering, University of Minnesota, has opened an office as consulting radio engineer with headquarters in the Munsey Building, Washington, D. C. Professor Jansky is a member of the Board of Direction of the Institute.

W. P. Koechel, until recently radio engineer, Westinghouse Lamp Company, Bloomfield, N. J., is now engineer, Ken-Rad Tube and Lamp Company, Owensboro, Kentucky.

B. H. J. Kynaston is now managing director, The Harmonic Radio Company, Ltd., Nottingham, England. Mr. Kynaston was formerly radio engineer of A. Holt and Company, Liverpool, England.

James McNary has recently resigned from the Radio Division, Department of Commerce, to become associated with Bell Telephone Laboratories in New York City.

Proceedings of the Institute of Radio Engineers
Volume 17, Number 12 *December, 1929*

GEOGRAPHICAL LOCATION OF MEMBERS ELECTED NOVEMBER 6, 1929

Transferred to the Fellow grade

Ohio	Cincinnati, 5449 Hamilton Ave.	Langley, Ralph H.

Transferred to the Member grade

California	Hollywood, 1041 N. Formosa Ave.	Hilliard, John K.
Dist. of Columbia	Washington, Radio Division, Dept. of Commerce	Terrell, William Dandridge
Illinois	Urbana, 301 Elec. Eng. Lab., University of Illinois	Brown, Hugh A.
Massachusetts	Longmeadow, 86 Dover Road	Forbes, Henry C.
New Jersey	Boonton, Fairview Ave.	Ferris, Malcolm
	Deal, Box 122	Ferrell, Enoch B.
	East Orange, 102 N. Arlington	Towner, Orrin W.
	Haddonfield, 19 Evergreen Lane	Diehl, William F.
New York	New York City, Western Union Tel. Co., 195 Broadway	Coggeshall, Ivan S.
	New York City, 40 E. 10th St.	Danziger, Harold I.
	Schenectady, Radio Eng. Dept., General Electric Co.	Loughren, Arthur V.
Ohio	Cleveland, Research Laboratories, National Carbon Co.	Nelson, James R.
Pennsylvania	Philadelphia, 570 W. Clapier St.	Deal, Harmon B.
	Wilkinsburg, 435 Ave. D	Beers, G. Lisle
England	South Croydon, 33 Dornton St.	Agate, C. S.
Holland	Eindhoven, 12 Jan Smitzlaan	van der Pol, Balth.

Elected to the Member grade

New York	New York City, Dept. of Commerce, Radio Service, Subtreasury Bldg.	Fish, Arthur S.
Philippine Islands	Manila, Fort Santiago, Radio Cable & Telegraph Office	Rives, Tom C.
Poland	Warsaw, Wspolna 32	Plebanski, Jozef
	Warsaw, Mokotowska 6, Radio Technical Institute	Sokolcow, Dmitri M.

Elected to the Associate grade

Alabama	Birmingham, 901 S. 38th St.	Duran, Albert E.
	Montgomery, 418 Adams Ave.	Persons, S. G.
Arizona	Prescott, O. A. Hesla Co., Box 658	deWitt, Harold
Arkansas	Little Rock, Y.M.C.A.	Tracy, Kermit F.
California	Los Angeles, 6220 La Mirada St.	Greger, J. G.
	Los Angeles, 405 E. Pico St.	Kruger, Bernard
	Los Angeles, 1467 Temple St.	Richards, W. S.
	Los Angeles, 5327 Monte Vista St.	Roberts, Russell B.
	Palo Alto, Federal Telegraph Co.	Lacabanne, Washington D.
	Palo Alto, 1040 Channing Ave.	Penther, Carl Joseph
	Palo Alto, 551 Lytton Ave.	Wagener, Winfield G.
	Redwood City, 523 King St.	Christensen, C. W.
	San Diego, 3000 A St.	Kinney, E. S., Jr.
	Tuolumne, Box 134	Elliott, William H.
Dist. of Columbia	Washington, c/o Bureau of Navigation, Navy Dept.	Pierrepont, John Jay
	Washington, Dept. Terres. Mag., Carnegie Inst. of Wash.	Seaton, Stuart L.
Illinois	Chicago, 1634 S. Springfield Ave.	Chernikoff, Leo J.
	Chicago, 6415 Stewart Ave.	Chatlos, Albert
	Chicago, 1925 Granville Ave.	Clark, Paul H.
	Chicago, 3710 Leclaire Ave.	Fitzgerald, John J.
	Chicago, c/o Shure Bros. Co., 335 W. Madison St.	Gram, Shirly L.
	Chicago, 3240 Evergreen Ave.	Grossman, Seymour M.
	Chicago, 5751 62nd St.	Hissong, Alfred
	Chicago, 122 S. Michigan Ave.	Lorch, George H.
	Chicago, 2611 N. Albany Ave.	Michalowicz, Leon V.
	Chicago, 809 Lill Ave.	Shanafelt, Lysle O.
	Chicago, Montgomery Ward & Co., Div. 62, Radio & Elect. Lab.	Thompson, Sidney J.
	Chicago, 4500 N. Sacramento Ave.	Waterman, S. S.
	Glenview, c/o WBBM	Fancy, Raymond C.
	Hinsdale, 206 Ayres Ave.	Doyle, E. J.
	Gary, 540 Lake St.	Gunther, W. J.
	Springfield, Sangamo Elec. Co.	Kunz, H. L.
Indiana	Valparaiso, Stiles Hall	Reynolds, Clay Elmer
Iowa	Cedar Rapids, 1718 C St. W.	Hajny, George F.
	Red Oak, 104 W. Reed St.	Anderson, Donald C.

Louisiana	New Orleans, 4515 Freret St.	Elliott, Harry M.
	New Orleans, 3421 Prytania St.	Dumestre, Alexis M.
	New Orleans, 1114 Pere Marquette Bldg.	Hancock, Olan W.
	New Orleans, c/o Western Union Tel. Co.	Hilgedick, W. C.
	New Orleans, 717 Greenwood St.	Scheffer, Roy J.
	New Orleans, 605 Belleville St.	Voegtlin, Elmo
Massachusetts	Chelsea, 124 Williams St.	Fox, Robert
	Marion, Radio Corporation of America	Brunette, Deo Z.
	Roxbury, 120 George St.	Ashenden, G. K., Jr.
	Springfield, 547 Page Blvd.	Cole, Neil D.
	Springfield, 32 Ardmore St.	Knapp, Harold D.
Michigan	Bay City, Radio WBCM	Carpenter, Ralph Harvey
	Detroit, 4322 Grand Ave. W.	Davis, Chester
	East Lansing, 348 Oakhill Ave.	Clark, Ralph L.
	Tecumseh	McConnell, Harley H.
Minnesota	Detroit Lakes	Hetland, L. C.
	Minneapolis, 3848 Harriet Ave. S.	Brooke, Robert O.
	Minneapolis, 915 Queen Ave. N.	Tynan, Thomas E.
Mississippi	Corinth, P. O. Box 541	Essary, William M.
New York	Auburn, 10½ Pleasant St.	Lumb, Frank J.
	Jackson Heights, L. I., 7517–41 Ave.	Fredendall, Beverly
	Long Island City, 3986 47th St.	Gilcher, V. J.
	New Brighton, 62 Westervelt Ave.	Macken, H. I.
	New York City, 961 Tiffany St.	Cilin, Louis
	New York City, Bell Telephone Labs., Inc., 463 West St.	Cole, Burton R.
	New York City, 195 Broadway, Room 2014	Herrmann, Henry J.
	New York City, 43 Exchange Place	Palmer, Robert T.
	New York City, 655 W. 160th St.	Stokvis, Morris, Jr.
	New York City, 1020 Walton Ave.	Weiland, Christian Frederick
	New York City, 281 E. 7th St.	Weiss, Samuel
	New York City, 615 W. 113th St. (Apt. 42)	Wilson, H. Warden
	New York City, c/o Technidyne Corp., 644 Broadway	Yolles, Jacob
	Schenectady, Research Lab., General Electric Co.	Nergaard, Leon S.
North Carolina	Asheville, Radio Station WWNC	Lance, Hubert H.
Ohio	Cincinnati, 3881 Reading Road	Klein, Helen
	Columbus, 625 W. 5th St.	Blum, Louis M.
	Lakewood, 17545 Madison Ave.	KaDell, Harold W.
Oklahoma	Tulsa, 1309 E. 15th St.	Banks, J. Vernon
	Tulsa, 2628 E. 6th St.	Carpenter, Hugh
	Tulsa, Radio Station KVOO	Richardson, Harry K.
Pennsylvania	Chester, 1219 Walnut St.	Mitchell, J. C.
	Cresson, R. D. #1	Vaughan, Kenneth A.
	Jenkintown, 309 Florence Ave.	Greenway, William L.
	Kane, 805 Welsh St.	Beatty, Rue Thompson
	Knox, Box 102	Smith, Jesse W.
	Philadelphia, 893 N. 48th St.	Chesny, John J.
	Philadelphia, 3422 Barclay St., E. Falls	Gerhard, Charles E.
	Philadelphia, 5220 Wayne Ave., 412 Qu Wayne Apts.	Morrow, Lorentz Arnold
	Philadelphia, 5531 Master St.	Stark, Harry W.
	Pittsburgh, 6503 Landview St.	Stayer, David
	Upper Darby, 202 Heather Road	Cahill, James A.
	Wilkinsburg, 414 Ella St.	Ballard, Randall C.
	Wilkinsburg, 815 Rebecca Ave.	Lehman, James N.
	Wilkinsburg, R. D. #1, Box 223	Sinnett, Chester M.
	Williamsport, 904 Railway St.	Petts, Ronald G.
South Carolina	Clemson College	Wilson, Walter B.
Tennessee	Nashville, 1918 Adelicia Ave.	Berry, Melvern H.
Texas	Beaumont, c/o KFDM, Magnolia Petroleum Co.	Bryan, William J.
Utah	Salt Lake City, c/o Radio KDYL	Barbre, Tom
Washington	Yakima, Tieton Drive	McQueen, Harry D.
Australia	Melbourne, University of Melbourne, Natural Phys. Lab.	Cherry, Richard O.
Canada	Montreal, P. Q., c/o Northern Electric Co., 637 Craig St.	Ketiladze, George S.
England	Enfield Wash, Middlesex, 22 Chestnut Road	May, John
	Hull, E. Yorks, Beverley Road, 42 Washington St.	Pereira, Francis Edward Duncan
	London, NW3, Hampstead, 10 Belsize Crescent	Goord, H. V.
	London, N7, Holloway, 46 Hilldrop Road	Huxter, Harold Charles
	Skipton, Dyneley House	Carr, John
	Watford, Herts, 31 Malden Road	Bartlett, A. C.
Holland	Hilversum, Nederlandsche Seintsestellen fabrien	Langendam, S. G. C.
India	Dharmsala Punjah, Govt. College	Vasudeva, D. N.
New Zealand	Christchurch, 180 Rolleston St.	Gibbs, R. J.
	Marton, Ngahima St.	Ruscoe, Chas. R.
	Wanganui, Post and Telegraph Dept.	Bradley, Ernest A.
Peru	Lima, Giron Camana 224	Maldonado, Arthur
Philippine Islands	Cebu, 37 Tres de Abril St.	Lauza, Nemesio D.

South Africa	Cape Province, Klipheuvel, Beam Wireless Station	Osborn, Eugene Wilson
	Johannesburg, 62 Persimmon Street Malvern	Lukat, John Frederick
South Wales	Cardiff, Llanishen, "Ciscar," Hilary Gardens, Cyn Coed Road	Jinman, Arthur Melville
Sweden	Stockholm, Kungsholmsgatan 21	Elmquist, Torsten

Elected to the Junior grade

California	San Francisco, Cal. Victor Dist., 536 Mission Ave	Lathrope, Kenneth W.
Indiana	Valparaiso, 712 Calumet Ave	DeHart, Delmar W.
Louisiana	New Orleans, 2434 Valence St	Dahlstrom, Hugo Wolf
Minnesota	Minneapolis, 3848 Harriet Ave. S	Brooke, Robert O.
Oklahoma	Alva, Radio KGFF, 709 Noble St	Sears, Garold D.
Canada	Montreal, P. Q., 637 Craig St., Room 808	Harvey, Fred E.

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APPLICATIONS FOR MEMBERSHIP

Applications for election or transfer to the various grades of membership in the Institute have been received from the persons listed below, and have been approved by the Committee on Admissions. Members objecting to election or transfer of any of these applicants should communicate with the Secretary on or before January 3, 1929. These applications will be considered by the Board of Direction at its January 8th meeting.

For Election to the Fellow grade

Germany	Berlin, Ludende, Karlstrasse 4	Fassbender, Heinrich

For Transfer to the Member grade

Massachusetts	Cambridge, General Radio Co., 30 State St.	Thiessen, Arthur E.
Pennsylvania	Irwin, 728 Grant St.	Marsteller, L. O.
Canada	Toronto, 45 Jarvis St.	Meredith, C. C.
France	Asnieres (Seine), 16 Rue de Chateaudun	Kraemer, G. I.

For Election to the Member grade

California	San Pedro, c/o Postmaster, *USS West Virginia*	Cooper, Lowell
Massachusetts	Springfield, 120 Ardmore St.	Lavallee, J. A.
	Springfield, 72 Princeton St.	Fortier, Ralph L.
New Jersey	Upper Montclair, 9 Bradford Parkway	DuMont, Allen B.
New York	New York City, c/o Radio Corporation of America, 233 Broadway	Grover, Harry G.
Ohio	Cleveland, Nela Park, Cleveland Vacuum Tube Works of G. E.	Ritter, E. W.
Pennsylvania	Pittsburgh, 5529 Bryant St.	Roth, Albert
	Wilkinsburg, 584 Lenox Ave.	Carlisle, Richard W.
Siam	Bangkok, Post and Telegraph Dept.	Prakit, Phya

For Election to the Associate grade

California	Berkeley, 1524 Scenic Ave.	Dempster, Everitt R.
	El Centro, 864 Olive St.	Irey, Earl R.
	Glendale, 1211 Winchester Ave.	Klingensmith, W. B.
	Hollywood, 4909 Sunset Boulevard	Brockway, Don C.
	Hollywood, 2144 Beachwood Drive	Heineman, John T.
	Los Angeles, 1129 4th Ave.	Brown, Cutler
	Los Angeles, 1023½ Blaine St.	Flickinger, J. H.
	Napa, P. O. Box 86	Streich, Robert J.
	Oakland, 1741 Seminary Ave.	Weaver, Chester E.
	Palo Alto, 151 Churchill Ave.	Boyson, Carl V.
	Palo Alto, 213 Fulton St.	Green, Irving L.
	Pasadena, 3209 E. Colorado St.	Craft, Walter C.
	San Francisco, 140 New Montgomery St.	Franklin, Lyman T.
	San Francisco, c/o San Francisco Examiner, Station KUP	Martin, Ronald G.
	San Jose, 87 E. San Antonio St.	Cooper, Charles A.
	Venice, 806 Marco Place	DuVall, Charles
Connecticut	Hartford, 52 Beacon St.	Grammer, George
	Naugatuck, Field St., R.F.D. #1	Wood, William Forrest
	Waterbury, 50 Hewlett St.	Buckley, Francis P.
Dist. of Columbia	Bellevue, Naval Research Laboratory	Kosar, William S.
	Takoma Park, P. O. Box 4253	Aurynger, John J.
	Washington, 911 Barr Bldg.	Codel, Martin
	Washington, 928 Madison St. N. W.	Dawson, William B.
	Washington, 226 8th St. S. E.	Guthrie, Robert C.
	Washington, 3333 P St. N. W.	Keough, Bayard Cole
	Washington, 3217 Brother's Place S. E.	Link, John C.
	Washington, 1319 F St. N. W.	Pierce, James F.
	Washington, 2807 Connecticut Ave.	Reynolds, George E.
	Washington, 214 17th St. N. E.	Shepherd, Henry Clayton
	Washington, 663 F St. N. E.	Taylor, Kenneth B.
	Washington, Navy Yard, Radio Test Shop	Tinsley, W. M.
Florida	Hialeah, c/o Tropical Radio Telegraph Co., Box 488	Corrigan, C. J.
	Tampa, 907 Florida Ave.	Weaver, Henry Ray
Illinois	Argo, 7806 W. 65 Place	O'Connor, James M.
	Chicago, 2543 S. Avers Ave.	Burger, Emil S.
	Chicago, 4614 N. Paulina Ave.	Elliott, John S.
	Chicago, 6800 Hilldale Ave.	Erwood, John
	Chicago, 656 W. Washington Boulevard	Huntley, Wilfred Eliot
	Chicago, 1305 S. St. Louis Ave.	Lyon, Alex

Illinois (cont'd)	Chicago, Radio Station WENR	McCluer, Paul
	Chicago, c/o Electrical Research Products Co., 910 S. Michigan Blvd	Manchester, Raymond Mark
	Chicago, 3333 N. Marshfield Ave	Menon, Geoffrey J.
	Chicago, 7712 Emerald Ave	Mower, Irving F.
	Chicago, 3942 Lake Park Ave	Rulison, Earl Wm.
	Chicago, Electrical Research Products Co., 910 S. Michigan Blvd	Sterling, M. F.
	Chicago, 4742 W. Adams St	Stokes, Howard S.
	Chicago, 1329 N. Wells St	Thineman, Edward H.
	Chicago, 2715 S. Komensky Ave	Tomas, Charles Frank, Jr.
	Elmhurst, 633 Spring Road	Walter, L. B.
	Glenview, Radio Station WBBM	Frisk, Caleb C.
	La Grange, 541 S. Kensington Ave	Beckley, John G.
	Plainfield	Barch, Henry Stefen
	Plainfield, Universal Wireless Comm. Co	Graham, Maxwell S.
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Part II

TECHNICAL PAPERS

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TYPICAL WIRELESS APPARATUS USED ON BRITISH AND EUROPEAN AIRWAYS*

By

EDWARD H. FURNIVAL

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Summary—*The information given in this paper is based in the main on the system in operation on British airways. European airways follow to a greater or less degree the general principles underlying the system described.*

SERVICE REQUIREMENTS

THE apparatus is designed to comply with the following service requirements:—

(a) Two-way communication is required, i.e., from ground to air and from air to ground.

(b) Reliable working is required normally up to distances of 100 to 150 miles, with the possibility of longer ranges being attainable up to a maximum of 250 miles when working with a main airport station.

(c) Radiotelephony is required for European airways. CW telegraphy, with the possibility of telephony, is employed on the Empire air routes, and also on certain European routes.

(d) Ground station receiving apparatus must be capable of giving directive reception so that bearings can be taken on transmission from aircraft, and by coordinated operation by a group of such stations, positions can be worked out by triangulation and passed to aircraft requiring information. Alternatively, a single bearing can be passed from one ground station alone.

(e) The wavelength on which communication is carried out between aircraft and ground stations is 900 meters.

(f) A separate hand-speed simplex telegraph service on a different wavelength is provided for inter-aerodrome traffic.

(g) A telegraphic broadcasting service is provided on another wavelength for the dissemination of meteorological information.

EXPLANATION OF SERVICE REQUIREMENTS

(a) Communication from aircraft to the ground stations is needed for the following main purposes:—

For distress calls. (This service is of primary importance to commercial aircraft flying to and from the Continent, owing to the channel crossing.)

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For reporting positions and progress.

For the acknowledgment of messages received from the ground stations

For position finding by the ground station direction-finders.

For reporting weather conditions of the upper atmosphere.

For announcing changes of course or destination.

For service messages, etc.

Communication from ground stations to aircraft is required for the following purposes:-

For passing weather reports.

For passing bearings and positions.

For giving information as to the proximity of other machines.

For acknowledging messages received from aircraft.

For passing service instructions, etc.

(b) As regards working range, this in the main is governed by the length of the stage or trip from one terminal to another. This distance may be of the order of 200 to 300 miles. Although intermediate aerodrome stations are available it is generally considered desirable to provide apparatus which will enable contact to be maintained by the aircraft with one or other of the main terminal stations, hence a reliable telephonic range of 100 to 150 miles is needed, telegraph ranges being proportionately greater.

(c) The most suitable type of transmission to employ has been the subject of much careful investigation. The reason for telephony is twofold. In the main it is to enable adequate communication to be carried out by flying personnel (either pilot or flying mechanic) without the intervention of the professional wireless operator. By this means encroachment on "pay load" is avoided, excepting by the carriage of the apparatus itself.

Incidentally it happens, however, that wireless telephony is particularly suitable in some respects for aerial communications. Given good intelligibility (by no means an easy attainment under flight conditions, but nevertheless one which can be obtained with well developed technique) communication is quick, and the pilot is enabled to maintain direct contact with the responsible official on the ground.

In the big machines and flying boats employed on the Empire air routes an operator is carried. Longer ranges are often needed, and it becomes advantageous to use CW telegraphy, with telephony as a stand-by, if required.

(d) It is a well-known fact that the directive properties of wireless can be used as a valuable aid to navigation. There are several ways which are capable of employment, but the one most generally used in Europe is by directive reception on the ground, and a network of such stations has grown up to serve the needs of commercial aviation.

It is usual to carry out normal communications on the direction-finding receiver, which can be immediately employed as a directional instrument when a call for a bearing or a position is received.

This system has the advantage that no extra weight or complication is involved so far as the aircraft is concerned, but where machines have to fly long distances, off the track of organized routes, then it becomes a case for carrying the direction-finding receiver on the plane, and special instruments have been developed for this purpose.

It is therefore normal to equip the modern airport with a direction-finding receiver of high grade. As is the case at the London Terminal Aerodrome, this may be situated actually in the control tower, the corresponding transmitting equipment being located outside the boundary of the aerodrome, remotely controlled from the control tower receiving station.

(e) For inter-aerodrome traffic separate transmitting and receiving equipment is installed to form an entirely independent channel of communication.

Messages concerning the arrival and departure of machines, service instructions, etc., are handled in this way.

(f) For the dissemination of meteorological information a separate wavelength is used, and the more important centers employ a separate station which sends out reports at stated intervals in the form of telegraphic broadcast messages in code. At present the following wavelengths are used:—

(a) For working between aircraft and ground stations 900 meters (this wave may be supplemented by others within the band 850 to 950 in the future, owing to traffic congestion).

(b) For inter-aerodrome working, 1400 meters. (214 kc per sec.)

(c) For meteorological services, 1260 meters. (238 kc per sec.)

(d) For emergency working between aircraft and ships and coast stations, 600 meters.

As it is possible that there will be a considerable revision in the allotment of wavelengths in the future, it was thought necessary to cope with this probable development by providing fairly generous wave ranges in all standard apparatus.

TECHNICAL DESCRIPTION OF APPARATUS—AIRCRAFT EQUIPMENT

A typical aircraft set which is commonly employed is the Marconi type AD6. This set is suitable for the transmission and reception of continuous wave, interrupted continuous wave, and telephone messages with aerodrome ground stations up to distances of 100 to 300 miles. Where space and personnel permit, the apparatus can be installed in such a position that the operator or observer has access to and full control over all adjustments, and can therefore take full advantage of the wave range and flexibility of the installation. In such cases, the dual control equipment is particularly useful, as it enables the pilot and the observer to communicate with one another without using the wireless.

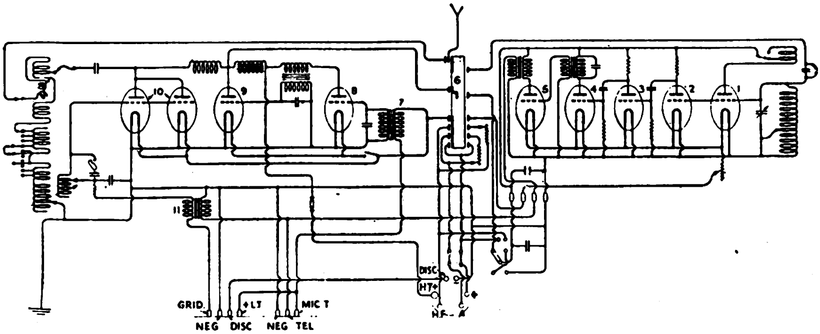


Fig. 1

(1) High-frequency amplifying valve.
(2) High-frequency amplifying valve.
(3) High-frequency amplifying valve.
(4) Detector valve.
(5) Low-frequency amplifying valve.
(6) "Send-receive" switch.
(7) Microphone transformer.
(8) Sub-modulating valve.
(9) Modulating valve.
(10) Oscillating valves.
(11) Side tone transformer.

Where space is restricted and where no observer or operator is carried, the set can be mounted in any convenient part of the machine, the remote control unit enabling the pilot to operate the set himself.

The wave range obtainable depends to a certain extent on the electrical capacity of the machine to which the set is fitted, which, in turn, depends largely on the size of the aircraft. When installed on a machine having an electrical capacity of about 0.0003 μf, the transmitter is adjustable to any wavelength between the limits 300 meters to 1,500 meters. The receiver is adjustable to any wavelength between 200 meters to 1,800 meters.

Owing to the extremely efficient radiator which is formed by a weighted wire suspended in space from the machine, this system has found universal application. Its advantages are twofold:— First, it enables the maximum range to be obtained from the minimum of power and weight, and secondly, such an aerial forms an extremely good "pickup" for the receiver, and this helps materially to maintain a favorable signal strength to ignition noise ratio, thereby tending to avoid ignition interference problems. Its disadvantage is the attendant risk of losing the aerial, especially during low flying under bad weather conditions, or by inadvertent loss when letting it out.

The modern tendency is to fit a supplementary fixed aerial on the larger aircraft, for low flying, especially in the vicinity of the aerodrome. The risk of loss due to operational causes is practically eliminated by the specially designed aerial winch which automatically controls the speed of release of the aerial.

The components of both transmitter and receiver are mounted in one box fitted externally with strong lugs, to which the rubber shock absorber suspensions can be conveniently attached. In spite of the extreme compactness of the apparatus, all components are readily accessible and can be removed by any qualified mechanic for examination or repair. A simplified diagram of connections is shown in Fig. 1.

The transmitting system comprises:—

(1) An aerial tuning inductance with tappings for varying the wavelength in steps and a variometer for obtaining intermediate values of wavelengths.

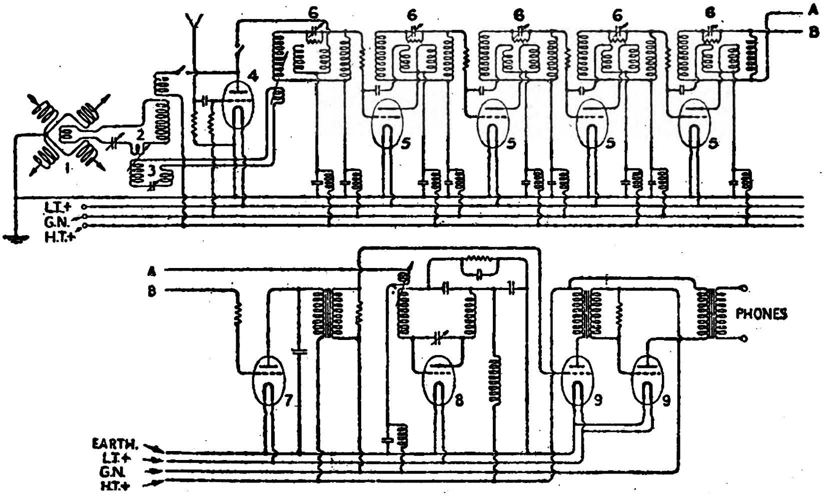
(2) Two oscillation valves for energizing the aerial circuit coupled to the aerial by means of a grid variable reaction coil and an anode tap connection on the aerial tuning inductance. A useful provision is the side tone circuit which enables the operator to overhear his own speech and to check whether the speech is being correctly transmitted.

(3) A modulating system consisting of a control valve and a sub-control valve connected in cascade across the generator and speech choke.

The receiving system comprises:—

(4) A tuner for high-frequency selection.

(5) A five-valve amplifier. One valve is used for reaction, two for high-frequency amplification, using "resistance-capacity" intervalve couplings, one for rectification and one for low-frequency amplification.


6
6
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5
L.T.+
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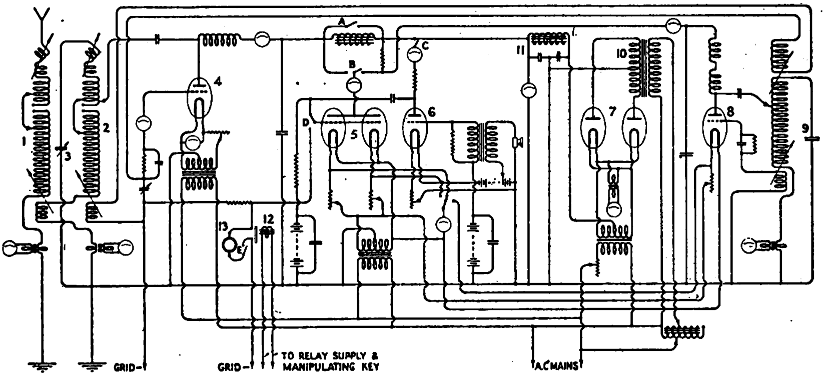
H.T.+
A
B
PHONES
7
8
9
9
EARTH.
L.T.+
G.N.
H.T.+

Fig. 2—Simplified diagram of connections of type R.G.14 receiver.

1. Directional selector.
2. Tuned search coil circuit.
3. Intermediate tuning circuit.
4. Open aerial coupling valve.
5. High-frequency amplifiers.
6. Tuned high-frequency grid circuits.
7. Rectifying valve.
8. Local oscillation generator.
9. Low-frequency amplifiers.


A
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10
11
12
13
GRID-
GRID-
TO RELAY SUPPLY & MANIPULATING KEY
A.C.MAINS

Fig. 3—Simplified diagram of connections of 4-kw type T.A.1 telegraph-telephone transmitter.

1. Aerial tuning inductance.
2. Closed circuit inductance.
3. Closed circuit condenser.
4. Power amplifier valve.
5. Modulating valves.
6. Sub-modulating valve.
7. Rectifying valves.
8. Independent drive valve.
9. Drive closed circuit.
10. Power transformer.
11. Smoothing system.
12. Signalling relay.
13. Interrupter disk.

Switches *A*, *B*, *C*, *D*, and *E* change the circuits to CW-ICW or telephony and are controlled by a single two-way lever.

The anode circuits of receiving valves are supplied with current from the generator. With the "send-receive" switch on the receive side the high-tension voltage of the generator is reduced to about 100 volts by the insertion of a high resistance in the high-voltage field of the generator.

The remote control unit is purely a mechanical device with four levers which operate, through Bowden cables, the following adjustments:—

(1) Change-over "send-receive" switch.
(2) Aerial tuning condenser.
(3) The receiver valve filament resistance.
(4) The receiver reaction coupling.

The unit is designed in a convenient form for mounting within easy reach of the pilot, and can readily be detached from the respective handles if and when required.

For the power supply a wind-driven generator is employed, operated by an automatic constant speed propeller. The generator used with the AD6 has a total output of about 180 watts, made up as follows:—

100 milliamperes at 1350 volts.
6 amperes at 7.5 volts.

The low-voltage output is sometimes increased and arranged to feed the lighting circuits of the aircraft through a suitable switchboard.

A small thin plate accumulator battery is connected so as to float across the low-voltage output. This arrangement is found very convenient for smoothing the ripple from the machine, and also for routine testing on the ground.

One or two other features are worth mentioning, namely:—

The microphone is of special construction, being highly damped to cause it to be relatively insensible to extraneous noise. The acoustic properties of the telephones are selected with care to ensure a quality of speech reproduction which can be readily understood above the surrounding noise. It is possible to shut out a considerable amount of noise by carefully designing the flying helmet, and it has been found desirable to issue a special helmet to ensure the most favorable results for reception, especially on noisy machines.

Intercommunication between pilot and operator or mechanic is arranged for by a separate attachment, while if telegraphy is required a keying unit with selector switch can be plugged into the main set.

The weight of the complete equipment is approximately 100 lbs.

Where greater power is required, such as in the flying boats operating in the Mediterranean section of the London to Karachi air route, the Marconi type AD8 installation is in use. The characteristics of this set are generally similar to the AD6, but the power input is increased to about 400 watts.

IGNITION INTERFERENCE

Where necessary this source of trouble is removed by the screening of the ignition system. Metal braided wire is used for the ignition leads associated with covers for the magnetos, and if necessary the spark plugs. Switch wires are often offenders, and require similar protection. Care has to be given to the application of this system, and it is desirable to make it as far as possible weatherproof.

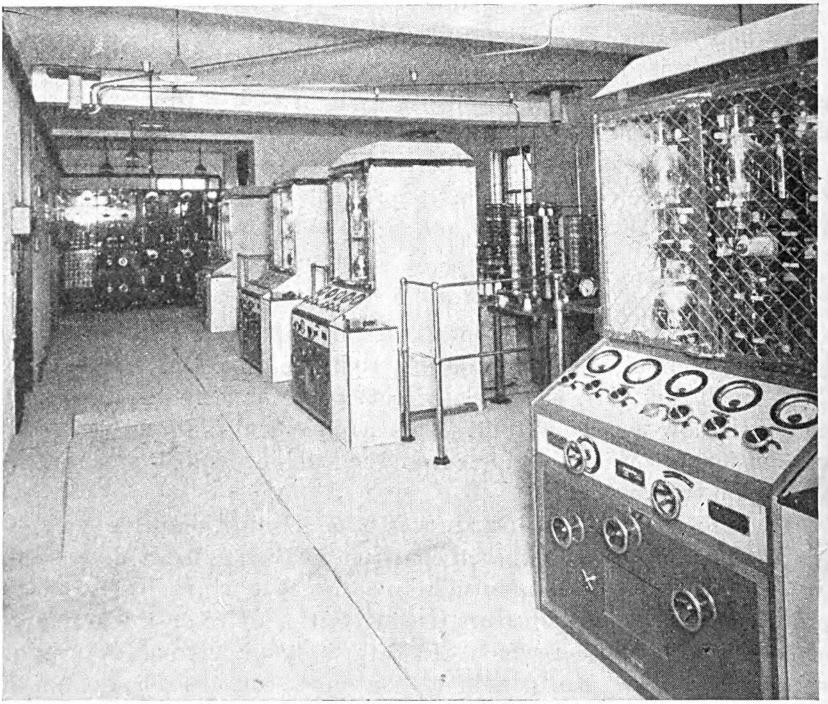

Fig. 4—Transmitting station at Croydon.

TESTING EQUIPMENT

In order to provide the simple routine procedure necessary for the effective testing and adjustment of aircraft wireless apparatus, the Marconi Company has developed a series of standard testing sets to permit of a series of accurate and precise tests. Their application enables an expert technician to carry out ground tests expeditiously and with the minimum of time and effort.

The principal item of this equipment which is in use at the terminal aerodrome is an instrument comprising a combination of an artificial aerial and a wavemeter. By the use of this device the equipment can be tested either *in situ* or on the bench. Oscillation condition, wavelength, and modulation characteristics of the transmitter can be readily checked while the receiver can be tested for sensitivity and wavelength with sufficient accuracy for practical purposes. For tests during flight a set of meters suitably mounted in unit form can be plugged into the set to test the power distribution to the filament, plate circuits, etc.

For tests on the ground a portable petrol engine is employed to provide power in place of the usual wind-driven generator.

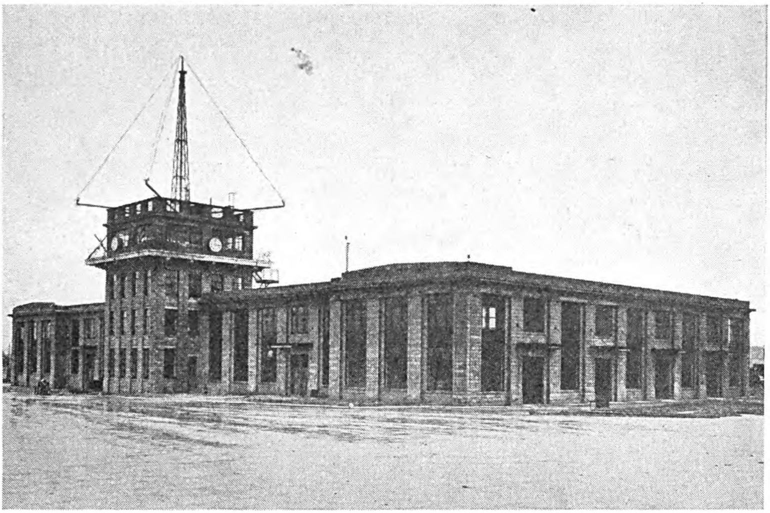

Fig. 5—Control tower showing direction-finding aerial system, Croydon.

By a carefully arranged system of ground tests aided by this equipment the maintenance engineers, assisted by reports from flying personnel and ground stations, are able to keep the equipment in an efficient state although they seldom have the opportunity of handling the apparatus in flight along the routes.

GROUND STATION EQUIPMENT

A typical modern ground station equipment for a main terminal aerodrome is the station recently established by the Air Ministry at the London airport at Croydon. This comprises a group of four transmitters working in conjunction with direction-finding receiving equipment.

In order to centralize and co-ordinate control, the direction-finding receiver is situated at the top of the control tower on the aerodrome. From here the operator, who works virtually side by side with the traffic officer, controls the transmitting equipment placed some 2½ miles from the boundary of the aerodrome.

Under the control of this operator the direction-finding receiver, on which all messages are received, was specially developed by the Marconi Company for this service, and is designed for the reception of CW, ICW, and telephone signals. A local oscillation generator is embodied in the receiver for the purpose of reception.

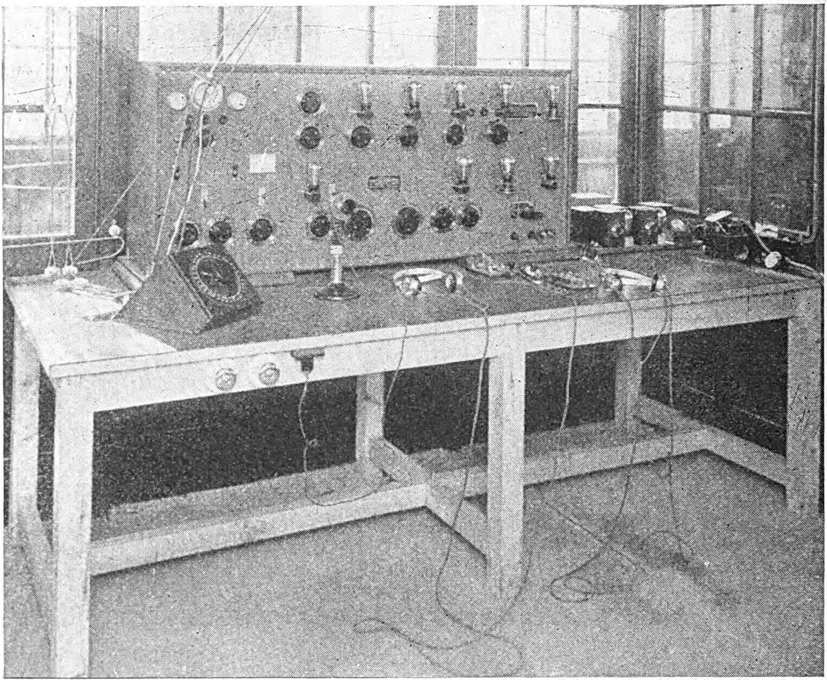

Fig. 6—RG14 installed in control tower at Croydon.

Three types of reception are provided for, namely, the "circle" or all-round diagram, the double directional or "figure-of-eight" diagram, and the uni-directional or "heart shape" diagram. A valve-coupled phasing unit incorporated in the receiver enables the uni-directional diagram to be made any desired shape, thus adding considerably to the selective properties of the receiver.

The receiver is used with two triangular fixed loop aerials and a single-wire open aerial, and radiogoniometers or directional selectors have been designed whereby more than one receiver can operate upon a common aerial system, thereby saving considerable expense when multi-way reception is required. The degree of accuracy obtainable with either the single or multi-way equipment is exceedingly high. For single-way reception the receiver is provided with a single direction-finding unit, and for multi-way working each receiver is provided with a double direction-finding unit, and a small artificial aerial unit.

The search coil of the radiogoniometer is tuned by a condenser, and can be coupled direct or via a tuned intermediate circuit to the special four-stage high-frequency amplifier. Each of the four high-frequency amplifying stages and the intermediate circuit is tuned by means of a condenser, and a gang control enables quick searching over a band of about ten per cent on either side of any desired spot wavelength to be obtained, at the same time keeping the overall amplification constant.

The various units of the receiver comprising the high-frequency tuning circuits, the high-frequency amplifier, the rectifiers, the local oscillation generator, and the low-frequency amplifier are mounted on the back of a vertical metal panel fixed to a rigid framework which slides into a teak containing case. The units are mounted in screening cases which can be easily removed when necessary. The various circuit adjustments and valves are mounted on the front of the panel.

With regard to the transmitting station, four separate and complete transmitters of the Marconi type T.A.1 are provided, each having a power input of 3 to 4 kw.

These sets are grouped together in one building, and are supplied with power from a motor alternator of sufficient capacity to feed the four units simultaneously. The motor alternator is duplicated, and a petrol electric plant is provided in reserve in case of failure of power from the supply mains.

Each transmitter has its own aerial system and each can be adjusted to operate on any wavelength between the limits of 800 and 2000 meters.

Provision is made in this type of set for the transmission of telephony and telegraphy either by CW or ICW, while an independent drive is incorporated to ensure a sufficient degree of wave constancy.

Modulation is by the well-known choke control method, the main control valves being arranged to act as absorber valves while keying.

The oscillation generator, modulating and rectifying system is contained in a metal cabinet, together with the independent drive apparatus, while external to this are mounted the astatically connected aerial tuning and closed circuit inductances and condensers.

A system of relays is provided to enable the plant (which is in charge of a station attendant) to be under the complete control of the aerodrome operator, the two stations being connected by a suitable multiwire underground cable for this purpose.

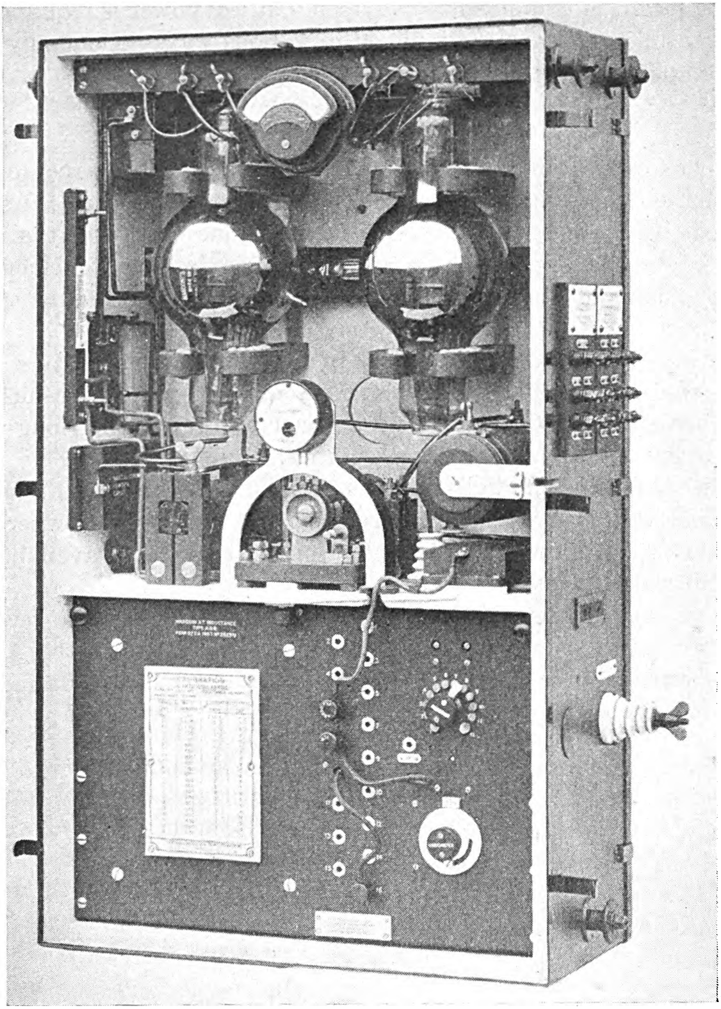

Fig. 8—Type AD8 instrument box.

Two of these transmitters are allocated to aircraft services; another is for the 1400-meter inter-aerodrome traffic; while the fourth is a stand-by set.

This plant has a guaranteed telephonic range of 250 miles to a standard aircraft set of the type A.D.6.

At intermediate airports similar telegraph-telephone and direction-finding receiving equipment is provided, but less power is required and the transmitting plant is simplified in various ways Generally only one complete transmitter is installed.

CONCLUSION

It has only been possible within the scope of this paper briefly to touch on the salient features of the apparatus described; furthermore, it has only been possible to mention one or two of the main types of stations employed. Constant expansion and improvement are taking place and what is current today will no doubt soon be superseded.

New methods and improvements in direction-finding are constantly being envisaged, while the employment of short-wave transmission, either as an alternative or as an adjunct to the existing long-wave system, has certain attractive possibilities.

The advent of the light plane movement also opens up rather a new field and special apparatus suitable for the private owner (the type AD.22) has therefore been developed and is now available for the light plane private owner.

A new type of set suitable for air-route working has also recently been developed (the AD.18). It is more powerful than the A.D.6, incorporates an independent drive, and has improved modulating and receiving circuits.

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RADIO FOR THE AIR TRANSPORT OPERATOR*

By
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Summary—*This paper deals with the aid of radio to an air transport operator; first, with merely long-wave receiving equipment on the plane, and, secondly, with the additional use of a short-wave plane transmitter. The tendency is noted toward establishment of short-wave two-way communication systems by the operators themselves. Requirements for apparatus under these conditions are discussed.*

ALTHOUGH air transportation has already demonstrated that it is an economic factor in modern life, it is equally true that radio in various types of application must play a large part in the program. This was wisely anticipated by Congress when they authorized the Department of Commerce to provide radio aids for air navigation. This work has been carried on admirably, and regular use is made of the present Airways Stations' weather and emergency broadcasts. Their radiobeacons, too, are helping to guide fliers through storm and night.

This is the simplest application of radio to the needs of air transportation in that the equipment on the airplane is light in weight, easy to install and operate, and least complicated in construction. To be sure, it does not offer the pilot communication to the ground, but it has a wide usefulness especially in single-man mail planes where the pilot is already seriously overworked in bad weather.

One of the oldest air mail carriers, National Air Transport, Inc., relies on the practical aid offered by this one-way radio service. This company has been experimenting with radio for almost two years and now has its Chicago-New York mail planes equipped with long-wave beacon and voice receivers, for routine operation. These are very sensitive instruments, using screen-grid tubes and are mounted in a compartment within the fuselage between the cockpit and the tail. Remote controls run up to the pilot so that tuning and adjustment of volume are very conveniently accomplished.

Some idea of the gain of the receiver may be indicated by the fact that 2-kw voice transmitters on 1000 meters are frequently picked up, in daylight, at distances of 300 to 400 miles using merely a 6-ft. mast as the antenna. This mast projects vertically from the fuselage about half-way between the cockpit and the tail of the plane. Much study and experimentation was necessary, of course, before the ignition interference from the engine could be reduced to negligible proportions.

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* Dewey decimal classification: R520. Original manuscript received by the Institute, September 12, 1929. Presented by M. M. Eells before New York meeting of the Institute, October 2, 1929.

Before "taking-off" with his mail cargo, the pilot studies the weather map and the latest reports available at the local Weather Bureau, concerning the actual conditions reported from observation stations along the airway. Certain regions along his line may look questionable, but he knows that subsequent weather reports will be broadcast regularly. However, he may request a special broadcast to be made at a specific time after his departure to give certain weather conditions as they may have developed in that particular questionable area.

Further than this, the one-way radio service has been of immeasurable value to National Air Transport, Inc., in a number of recent instances where the terminal fields have been closed in by fog. Emergency broadcasting by the Airways Radio Station was arranged with the assistance of the local field manager who could hear the plane as it roared through the fog near the field. By personal radio instructions to this pilot it was possible to bring him back and forth across the field until he could be brought down gradually in safety.

The fund of experience National Air Transport, Inc., has gained in flying these radio-equipped planes has proved that the ground-to-plane radio aids are of tremendous value in bad weather flying, with assistance offered in the four following degrees:

(1) Reduction in flying time by more direct navigation.
(2) Flights made that would not otherwise have been attempted.
(3) Prevention of pilots becoming lost and having to leave plane.
(4) Avoidance of crashes and fatalities in thick weather.

To consider the completion of the communication setup, it is obvious that much additional assistance can be given the pilot if he is able to talk with the ground. With these facilities, he can materially aid the Weather Stations in their collection of information, by reporting conditions he meets in flight. He is able to ask for specific information which may be of most vital importance if he is shut in by fog or storms. In case of a forced landing, the operation of the transmitter even if in only an inefficient emergency manner, may mean everything in assistance to those trying to help him.

Ground receiving equipment can be maintained by the operating personnel at the fields, but there is the question of getting the reply back to the plane. Where there is a local Department of Commerce voice broadcaster, the message can be handled by them as an emergency broadcast, if the reply is of the nature of life, property, or cargo.

It is recognized, however, that the large air transport operators will prefer ultimately to have two-way communication systems of their own. Short waves, below 200 meters, will be used for both ground-to-plane and plane-to-ground. Ground stations will be owned and operated individually by the air transport operators or cooperatively if the routes of two or more operators join or are parallel. Such a system of privately operated radio facilities will provide means for dispatching planes, communication at times of distress for safety of life and property, and means of handling commercial messages to and from passengers in planes at such a time when this may be an economic step. Course indication or triangulation may be made where direction-finding ground equipment is provided. Pan American Airways and Transcontinental Air Transport have been employing this means as an aid to navigation.

Telephone is the preferred form of communication especially for single-pilot planes on account of greater speed message-sending and because of the simplification of the operator problem. Certain of the larger transport planes are, however, using code, especially where greater distances must be covered or where static is unusually heavy. This means a saving in equipment weight against which is balanced the extra radio operator.

The question of communication between airports is naturally a very vital one, not only for plane dispatching but also for weather dissemination. Interfield communication should not be conducted by radio where existing land lines can handle the traffic without hazardous delay. Such a practice, if adopted generally, would clutter frequency channels sorely needed for legitimate services that can use only radio.

The air transport operator has a difficult problem in finding suitable equipment. A good portion of this is due to the comparative lack of experimental data on short-wave airplane operation. Ordinary commercial equipment is not suitable until it has been thoroughly redesigned from the standpoint of weight, size, extreme durability, and reliability. Cushioning of the parts against vibration in flight, or jars due to heavy landing, presents a problem in itself.

The balancing of weight against reliability of operation must always be a compromise. When it is realized that failure of the installation in flight may mean loss of life and property, too much attention cannot be given to each slightest detail.

However, it must not be forgotten that the air transport operator pays dearly for each extra pound of weight carried. It is not impossible to load the airplane with so much extra material in the interest of safety and reliability that there is no room left for economic pay load.

Two of the most difficult problems are the source of power and the antenna for the transmitter. If the plane is to be able to transmit from a forced landing, a dynamotor driven from the plane's 12-volt battery is preferred to an engine-driven or wind-driven generator. A heavy current is drawn, in the neighborhood of 55 amperes, which, with the 70-ampere landing lights, makes a lot of work for the charging generator. There seems tobe good promise in a double-voltage engine-driven generator where a disconnecting clutch can be incorporated to allow the generator to be driven as a dynamotor in emergencies. Wind-driven generators are easy to install and their cooling problems are somewhat simplified, but they make testing and ground operation a very complicated matter. Their wind resistance may be enough to raise the equivalent weight to such a point that the air transport operator loses interest.

The antenna, of course, depends upon the frequency used. Again a compromise comes up as to whether the operator wishes the increased operating range obtainable with a trailing wire or a shorter working distance in order to have a fixed antenna structure. In either event caution must be exercised so that no maneuver or breakage of the antenna can result in fouling control cables. If the fixed antenna is definitely required, it appears that frequencies above 3000 kc must be used. To give adequate day operating range, the operator will probably have to change frequencies, which seriously complicates the maintenance problem. It may be necessary to answer some of these questions by locating ground transmitters more closely together.

There is no question as to the assistance radio can give to the air transport pilot. The problem is to insure against failure of the system. A catastrophe may be easily caused if the pilot relies on radio at a future instant and it fails him for one cause or another. It would be far better to have had no radio at all. The simplest installation of today is far too complicated and too fragile. Some of the best technical brains in the country are studying this vital phase of reliability, but it is a far-reaching complication.

The one realization remains that radio is the only means of contacting with airplanes in flight; that a communication system is vital to the economic development of this wonderful means of transportation; and that this all-important need will eventually overcome obstacles that now may seem insurmountable.

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THE CIVIL AIRWAYS AND THEIR RADIO FACILITIES*

By
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Summary—*This paper describes the establishment of the radio facilities on the civil airways of the United States and their importance for air navigation. The legislation which created the Aeronautics Branch of the Department of Commerce and its organization is discussed.*

Description is given of the radiotelephone and radiotelegraph equipment used in the collection and broadcasting of weather information to aircraft. The transmitter used is rated at 2-kw output and is of the master-oscillator controlled type.

A directive radio transmitting system using crossed coil antennas is employed to guide aircraft over the established airways. It is of such a type that the pilot receives one signal while within a narrow zone on the course, and other signals if off the course. This greatly facilitates flying under conditions of poor visibility.

The regulations that have been proposed for the installation of radio equipment on aircraft are briefly given together with the more important frequencies allocated for aircraft work. The purpose and needs for these radio facilities and the way in which they have developed are also discussed.

THE Air Commerce Act, approved May 20, 1926, provided for the promotion and regulation of civil aeronautics by the Department of Commerce.

The Act authorized the establishment and maintenance of civil airways equipped with intermediate landing fields, beacon lights, signal and radio apparatus, and other aids to air navigation, including facilities for weather-reporting service; the promulgation and enforcement of rules and regulations to control traffic in the air; the marking of all civil aircraft by identification number and the regular inspection and licensing of commercial aircraft engaged in interstate flying; the examination and licensing of airmen, i.e., pilots and mechanicians; the charting of airways and the publication of air maps; the dissemination of information about air commerce, especially including data concerning accidents and their causes; scientific research and experimentation designed to improve the facilities for air navigation and transport; and, generally, the promotion of air commerce, industry, and trade.

To carry out the above purposes the Aeronautics Branch of the Department of Commerce was organized. This Branch is a part of the Office of the Secretary of Commerce and embraces services of several Bureaus of the Department. The following divisions are included in the Aeronautics Branch: Division of Air Regulations, which examines and licenses pilots and mechanicians, inspects aircraft to determine their air-worthiness, enforces the air traffic rules and studies airplane engines and designs; the Division of Air Information, which collects and disseminates information about civil aeronautics; the Airways Division (a part of the Lighthouse Service), which establishes and maintains aids to navigation along commercial air routes; the Airways Mapping Section, (a part of the Coast and Geodetic Survey) which produces the special maps needed; and the Aeronautical Research Division, (a part of the Bureau of Standards) for conducting research which would tend to improve the facilities of air navigation.

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* Dewey decimal classification: R520. Original manuscript received by the Institute, September 16, 1929. Presented before New York meeting of the Institute, October 2, 1929.

There are at present in the United States approximately 52 air transport companies flying regular schedules covering approximately 40,000 miles per day. Twenty-seven of these companies carry mail over the lighted airways, and about 15,000 miles of the flying takes place at night.

A double mail schedule is now in operation on the Transcontinental Airway, which saves a business day in each direction as compared with the former schedule. Mail gathered at the close of one business day on the Pacific coast will be delivered 36 hours later in cities on the Atlantic seaboard, the mail being in transit two nights and one day. Two transport companies have established a coast-to-coast passenger service and many other companies are carrying passengers over lesser distances.

The success of these services will depend very largely upon safety, regularity of flying, and percentage of trips completed on scheduled time. This can be assured only by the use of radio, a comprehensive weather service, and improved instruments aboard airplanes, making it possible to fly safely with regularity regardless of weather conditions and visibility. With proper facilities a schedule efficiency of 98 per cent should be possible.

Radio, by its fundamental nature, can provide for air navigation two very important services, viz., communication between the aircraft and the ground, and directional indications. The value of such services has been well demonstrated in marine navigation but, it is believed, they will be of even greater value to aircraft. The value of radio in this connection was appreciated by Congress and the establishment of radio systems was included in the Air Commerce Act.

When plans for the establishment of an aeronautical radio communication system were being formulated there was little experience in this country that might be followed. European practice, however, favored the use of frequencies in the 315–350-kc band for this purpose.

At about this time also, preparations were being made for the International Radiotelegraph Conference to be held in Washington in October, 1927, and the European countries had proposed that the 315–350-kc band be internationally used for aeronautical services. With a need for such a service in the United States, the American delegation agreed to this arrangement, and these frequencies were adopted internationally for this service. International agreement was also secured on the band 285–315 kc for radiobeacons for both the marine and aircraft services.

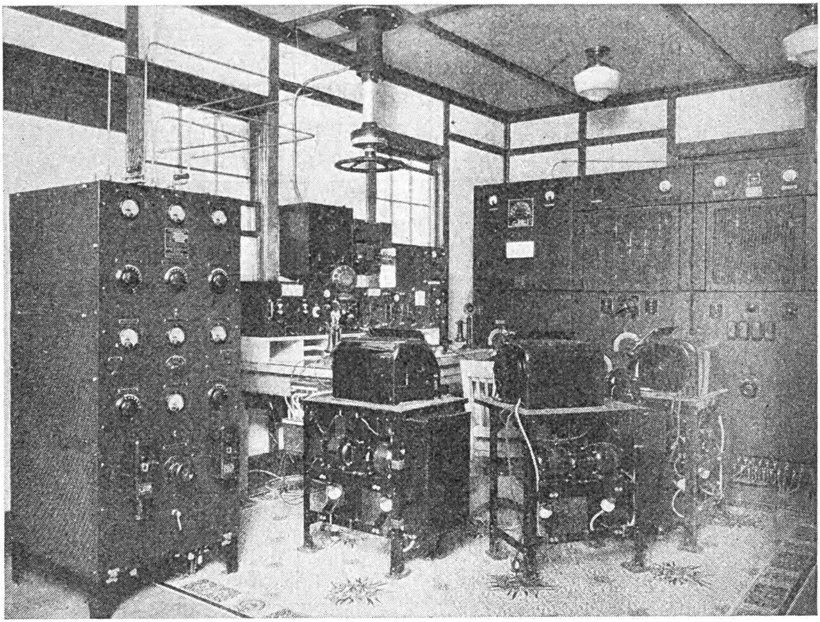

Fig. 1—Interior of aeronautical radio station at Glendale, Calif.

In providing a radio service for the transmission of weather and safety messages to pilots in the air it was considered that radiotelephony would be most practicable. For this class of service, a combined telephone and telegraph transmitter was believed most desirable, since it would provide not only for the broadcasting of weather information to aircraft, but also for the collection of such information by radio from similar radio stations. In the selection of this equipment, the experience of the Navy and the Army was carefully considered. A type of transmitter was finally adopted that was generally similar to one that had been widely used by the Navy, both afloat and ashore. This transmitter is of the master-oscillator, intermediate-amplifier, power-amplifier type, and will deliver 2 kw of radio-frequency power to an antenna. Either CW, ICW, or telephone operation is provided. The transmitter will operate on frequencies from 100 to 550 kc. This transmitter is shown on the right of Fig. 1. A 400-watt crystal-controlled high-frequency telegraph transmitter is shown on the left. This transmitter is also used in collecting weather information as mentioned above, but is not capable of telephone operation. The performance of the intermediate-frequency transmitter has in general been satisfactory. Harmonic radiation has been excessive, but this is being reduced by working the power amplifier into a tuned tank circuit with capacitive antenna coupling instead of an untuned transformer. A single-wire antenna 125 ft. high and 375 ft. long is used with this transmitter.

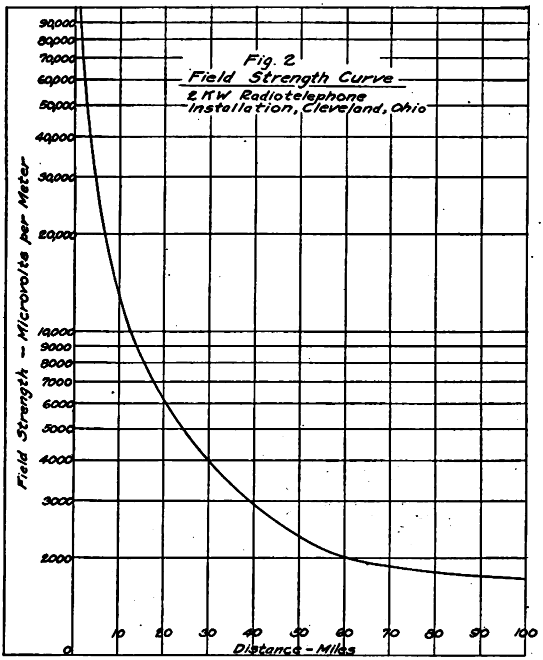

Fig. 2—Field-strength curve. 2-kw radiotelephone installation, Cleveland, Ohio.

The field strength [1] produced by this transmitter as measured on the ground is shown in Fig. 2. No field-strength measurements have been made in the air although this is contemplated as soon as a plane can be made available. The field-strength curve has an attenuation of approximately 0.0062, which is somewhat lower than found on frequencies in the broadcast band. Telephone transmissions are regularly received at ground receiving stations at a distance of 200 miles in daytime through ordinary static. When received on a airplane equipped with commercial receivers, the reliable range is approximately 150 miles under ordinary conditions and approximately 60 miles under severe conditions. An improvement in performance is expected with changes and improvements in receiver design, reduction in motor and air noises, sound-proofing of cabins, better bonding of the plane's members, and better ignition shielding.

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[1] These measurements were made on an installation at Cleveland, Ohio, by the Supervisor of Radio of Detroit, Mich.

Radio stations as described above have been located at intervals of about 200 miles along the airways. This spacing was determined partly by the estimated reliable range of this power transmitter and partly because a given mail plane usually covers a 400-mile section with a stop midway for gas. Large cities served by the air mail service also appear at approximately 400-mile intervals. The first transmitter of this type was installed at Hadley Field, New Brunswick, N. J., in December, 1927, and since that time twenty-four additional stations have been completed and placed in operation. The map (Fig. 3) shows the installation program as of June 30, 1929. Stations now in operation are located at New Brunswick, N. J.; Bellefonte, Pa.; Cleveland, Ohio; Bryan, Ohio; Maywood, (Chicago), Ill.; Iowa City, Iowa; Bellevue, (Omaha), Nebr.; North Platte, Nebr.; Cheyenne, Wyo.; Rock Springs, Wyo.; Salt Lake City, Utah; Elko, Nevada; Reno, Nevada; Oakland, Calif.; Fresno, Calif.; Glendale, (Los Angeles), Calif.; Medford, Ore.; Portland, Ore.; Seattle, Wash.; Anglum, (St. Louis), Mo.; Kansas City, Mo.; Wichita, Kans.; La Crosse, Wis.; and Key West, Fla.

Equipment has been purchased for additional stations at Albany, N. Y.; Atlanta, Ga.; Birmingham, Ala.; Boise, Idaho; Boston, Mass.; Buffalo, N. Y.; Fort Worth, Tex.; Greensboro, N. C.; Houston, Tex.; Jacksonville, Fla.; Miami, Fla.; Milford, Utah; Mobile, Ala.; Nashville, Tenn.; New Orleans, La.; Oklahoma City, Okla.; Pasco, Wash.; Richmond, Va.; Spartanburg, S. C.; and Washington, D. C.

The establishment of directive radio facilities, however, was more of a problem. The U. S. Lighthouse Service had had considerable experience with the non-directive type of radiobeacon[2] in its Marine Service, using the system of radio direction-finding in which the transmitting equipment is located on shore and radiocompass is located on the ship. This system has been in operation since 1921

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[2] G. R. Putnam, "Radio fog signals and their use in navigation with the radio compass." U. S. Bureau of Lighthouses, 1924.

with radiobeacons installed at important lighthouses and on lightships. Many bearings on these installations are being taken every day by masters of vessels by means of the radiocompass aboard their ships. This system has many advantages, but did not seem applicable to present day aircraft. It is believed, however, that with larger aircraft such a system may yet come in general use. The radiobeacons required for such a system are as a matter of fact provided at the present time, since the radiotelephone equipment described above is operated on definite schedules and will provide, with its circular radiation characteristic, a suitable signal on which bearings may be taken with a radiocompass. With one-man planes in which the pilot is navigator, engineer, captain, and radio operator combined, it did not seem possible for him to take bearings with a radiocompass and it was considered undesirable to use a fixed loop on the plane and take bearings by changing the course of the plane. This eliminated the nondirective type of radiobeacon. Consideration was also given the revolving radiobeacon,[3] which requires an ordinary receiver and a stop watch, but this was considered unsatisfactory also since it required some manipulation on the part of the pilot. The so-called "aural" type of directive radiobeacon, or "radio range" as it is now called, was

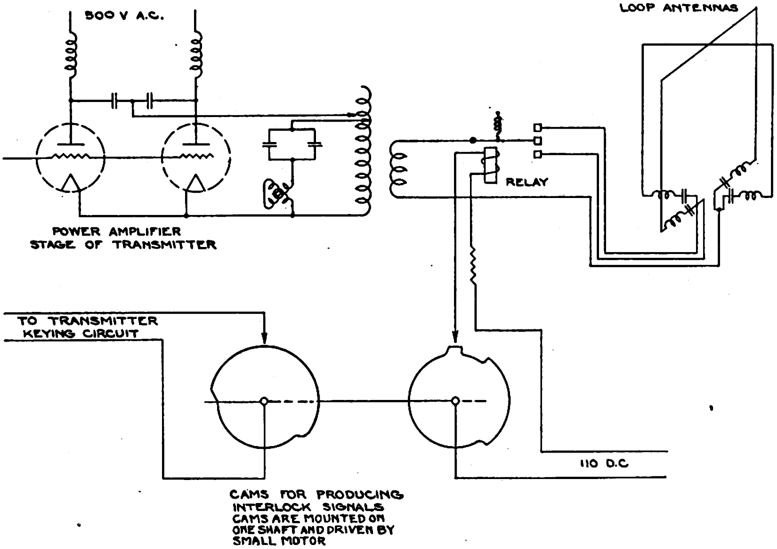
500 V A.C.
LOOP ANTENNAS
RELAY
POWER AMPLIFIER STAGE OF TRANSMITTER
TO TRANSMITTER KEYING CIRCUIT
110 D.C.
CAMS FOR PRODUCING INTERLOCK SIGNALS CAMS ARE MOUNTED ON ONE SHAFT AND DRIVEN BY SMALL MOTOR

Fig. 4—Essential circuits of a radio range.

[3] Smith-Rose and Chapman, *Jour.* I. E. E. (London), 66, 256–69; March, 1928.

finally considered to be most applicable to the Airways Service. Some preliminary experiments with this type of directive radiobeacon had been made in 1921 by several departments of the Government in cooperation with the Lighthouse Service, in which observations were made by the Lighthouse Tender "Maple" in the Potomac River and Chesapeake Bay, and also by airplanes,[4] on signals transmitted from Washington. Further improvements on this type of directive radiobeacon were made by the Signal Corps of the U. S. Army.[5]

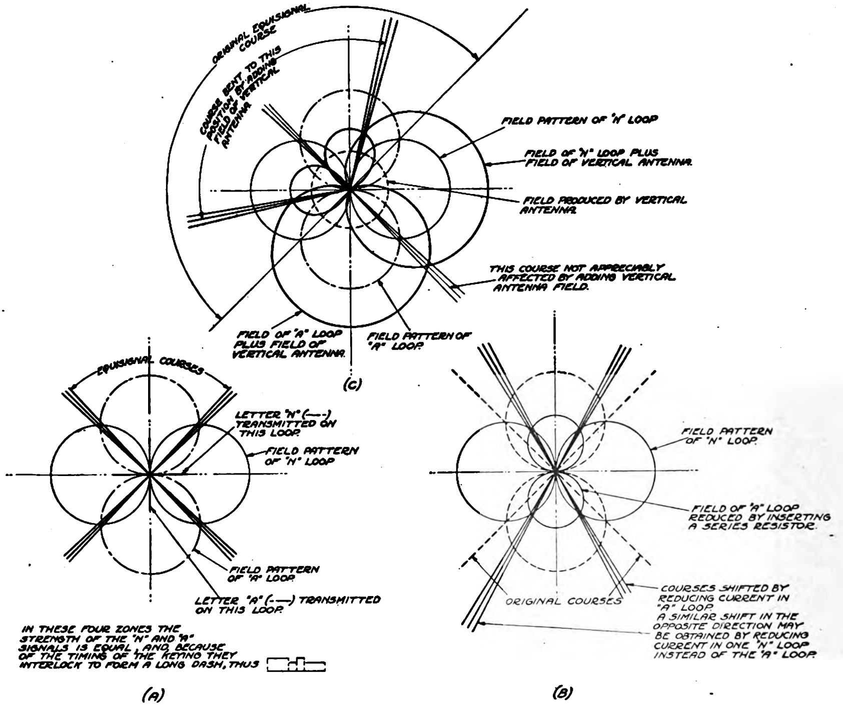
ORIGINAL EQUISIGNAL COURSE
COURSE BENT TO THIS POSITION BY ADDING FIELD OF VERTICAL ANTENNA
FIELD PATTERN OF "N" LOOP
FIELD OF "N" LOOP PLUS FIELD OF VERTICAL ANTENNA.
FIELD PRODUCED BY VERTICAL ANTENNA
THIS COURSE NOT APPRECIABLY AFFECTED BY ADDING VERTICAL ANTENNA FIELD.
FIELD OF "A" LOOP PLUS FIELD OF VERTICAL ANTENNA.
FIELD PATTERN OF "A" LOOP.
(C)
EQUISIGNAL COURSES
LETTER "N" (—-) TRANSMITTED ON THIS LOOP.
FIELD PATTERN OF "N" LOOP
FIELD PATTERN OF "A" LOOP.
LETTER "A" (-—) TRANSMITTED ON THIS LOOP.
IN THESE FOUR ZONES THE STRENGTH OF THE "N" AND "A" SIGNALS IS EQUAL, AND BECAUSE OF THE TIMING OF THE KEYING THEY INTERLOCK TO FORM A LONG DASH, THUS
(A)
FIELD PATTERN OF "N" LOOP.
FIELD OF "A" LOOP REDUCED BY INSERTING A SERIES RESISTOR.
ORIGINAL COURSES
COURSES SHIFTED BY REDUCING CURRENT IN "A" LOOP. A SIMILAR SHIFT IN THE OPPOSITE DIRECTION MAY BE OBTAINED BY REDUCING CURRENT IN ONE "N" LOOP INSTEAD OF THE "A" LOOP.
(B)

Fig. 5—Field patterns of radio range transmitter.

The transmitter used by the Airways Division in its radio range installations is of the master-oscillator, intermediate-amplifier, power-amplifier type, with special precautions taken to maintain constant frequency. Transmitters with an output of about 1500 watts are now used which give a reliable service up to approximately 100 miles. The field strength produced is much less than that given in Fig. 2, since loop antennas with their poor radiation characteristic are required.

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[4] Engel and Dunmore, "A directive type of radio beacon and its application to navigation," Bureau of Standards Scientific Paper No. 480, 1923.

[5] Murphy and Wolfe, *Jour.* S. A. E., 19, 209; September, 1926.

Briefly, the radio range consists of the transmitter mentioned above, working into two loop antennas placed at 90 deg. to each other, with a relay that connects the radio-frequency power alternately to the two loops. The essential parts of the radio range are shown in Fig. 4, and the radiation characteristics produced are shown in Fig. 5. The transmitter is keyed so as to produce long dashes, and the relay in the output circuit breaks up the dash so that the letter *N* (dash dot) is transmitted on one loop and the letter *A* (dot dash) on the second loop. By examination of the radiation characteristic, Fig. 5(a), it is noted that there are four zones each 90 deg. apart in which the signals from both loops have the same strength, and the letters *N* and *A* add together or interlock to produce the long dash of the transmitter keying. The equisignal zones may be directed along the airways to be served by

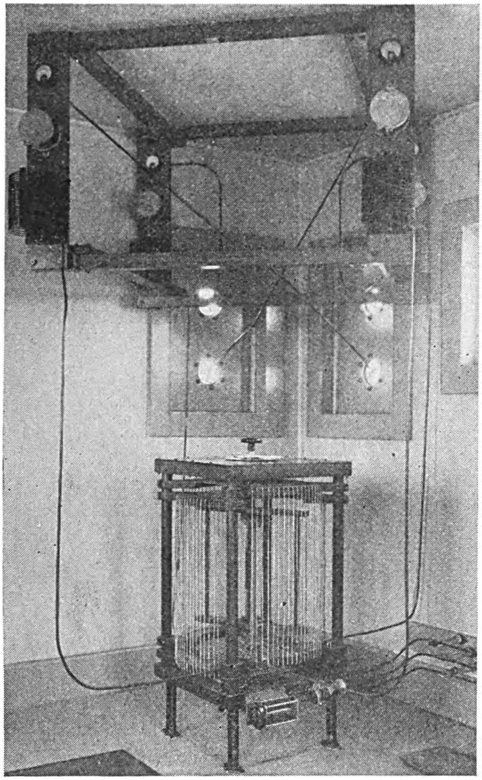

Fig. 6—Goniometer and loop tuning equipment originally installed at Hadley Field.

proper orientation of the loops. A pilot, in flying the airway, would hear a long dash when he was on the course and should he drift to the *A* side the *A* signal would increase and become predominant while the *N* would fade to the background and vice versa.

In order to direct the courses more easily along the route desired and to facilitate installation, use is made of a "goniometer" similar to that used in the Bellini-Tosi direction-finder.

This type of directive radio transmission appeared to meet the problem best of all, since radio courses can be established along given airways which would carry the pilot on a straight line over the best flying country along which weather observations were made, and intermediate landing fields established.

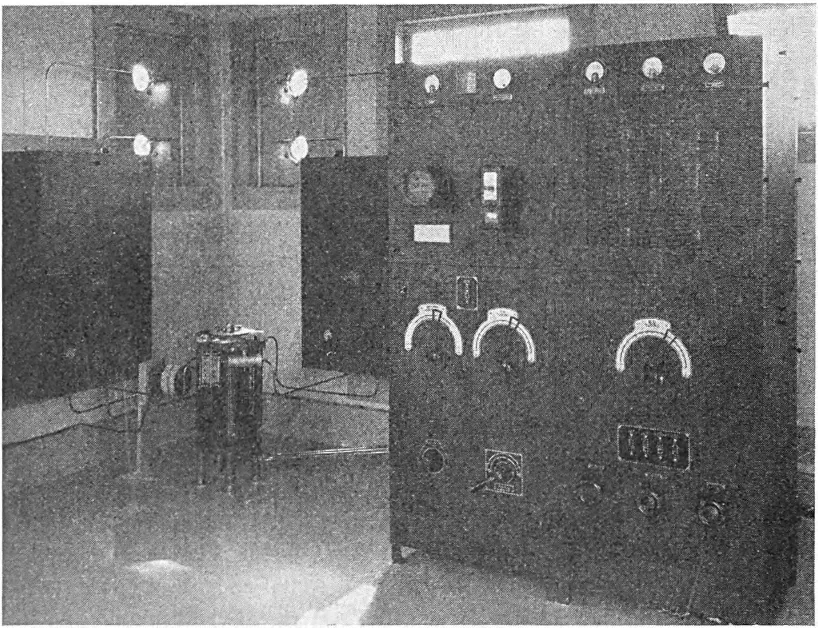

Fig. 7—Radio range installation at Hadley Field.

This type of equipment also has the important advantage of eliminating drift errors caused by a cross wind. Another important advantage is that the sense of hearing is employed rather than the sense of sight, since in bad weather flying the pilot must of necessity be watching the ground carefully to check up on landmarks. In blind flying, this system is equally advantageous since the pilot must concentrate on certain instruments to keep his ship on an even keel.

In order to gain in experience with this directive radio system a complete installation was made in cooperation with the General Electric Co., and this equipment was installed at Hadley Field near New Brunswick, N. J., in June, 1927. Tests made on this installation extending over a period of about seven months indicated the need for further improvements but confirmed the belief that this system was adequate for the purpose intended. Utilizing the experience gained, specifications were prepared covering the improved apparatus desired, and six complete transmitters of this type were purchased. Fig. 6 shows the original goniometer and loop tuning equipment installed at Hadley Field. Fig. 7 shows the improved transmitter which replaced the test equipment. Fig. 8 shows the radio range building and loop antennas.

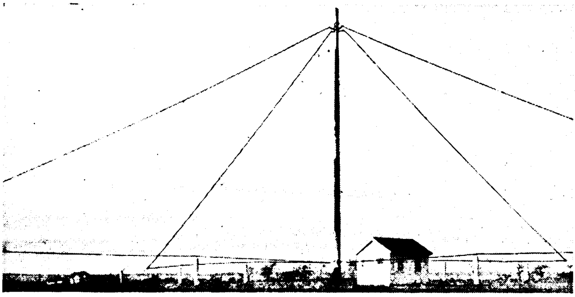

Fig. 8—Exterior of Hadley Field radio range.

A glance at Fig. 9, a map issued by the Department of Commerce, showing the civil airways of the United States, reveals the fact that airways are not always straight lines nor do they radiate from the larger cities with any mathematical relation. To adapt the directive radio system described above to these airways required that methods be developed whereby the four courses set up by the loop antennas could be shifted or "bent" as required to line up with the already established airways. The engineers of the Airways Division developed two methods whereby this can be effected. Fig. 5(a) gives the normal radiation characteristic of the radio range. Four courses are provided at 90 deg. apart since the radiation of each loop is the same. The first method of shifting the courses is by increasing or decreasing the current in either loop as by removing or inserting resistance. The effect of this is shown in Fig. 5(b). The second method of shifting or "bending" is accomplished by adding the circular radiation from an open vertical antenna to that of the loops. This produces courses "bent" as shown in Fig. 5(c), and the degree of "bending" is proportional to the current in the vertical antenna. In some installations an inherent bend has been noted which it is believed is caused by the circular radiation set up by currents induced in an iron conduit carrying an obstruction light circuit to the top of the center antenna pole.

Radio ranges are now in operation at the following locations: Hadley Field, New Brunswick, N. J.; Bellefonte, Pa.; Cleveland, Ohio; Goshen, Ind.; Sterling, Ill.; and Des Moines, Ia. Two additional installations are being completed at Boston, Mass., and Chicago, Ill. It is now possible to fly from Hartford, Conn., to Omaha, Nebr., utilizing directive radio signals from these installations, without ground observations that would be otherwise required. This is of much importance at times of poor visibility. Fig. 10 shows the courses marked by these installations and it will be noted that the courses from practically all installations required some "bending" of the courses to serve the cities along the airway.

Adjacent radio ranges are operated on different frequencies, and "marker beacons" of short transmitting range are located at the intersection of adjacent courses (designated as *M* in Fig. 10). The marker beacons send a special characteristic signal modulated at about 1000 cycles, alternately on the frequencies of the intersecting courses, thereby informing the pilot that the intersection has been reached and that his receiver should be re-tuned to the frequency on which the adjacent radio range is operating. Marker beacons operating on a single frequency are also located at high points on the airway and at other points to mark particular obstructions. These marker beacon transmitters have a maximum output of about 5 watts, and while it is desirable to have their maximum effect (i.e., transmitting range) not more than five miles, difficulty is experienced in keeping them within this range.

This system, like all other radio systems, is not free from peculiar phenomena of wave transmission. Particularly at night, shifting of the equisignal zone is observed.[6] This shifting has not appeared to be a serious matter when a vertical pole antenna is used, for, by taking the on-course and off-course significance of signals over a short interval the average is approximately along the proper course, and over distances of less than the hundred miles which are here involved, the shifting is of small magnitude.

There has been considerable research on a type of radio range that

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[6] H. Pratt, "Apparent night variations with crossed coil radio beacons," Proc. I. R. E. **16**, 652; May, 1928.

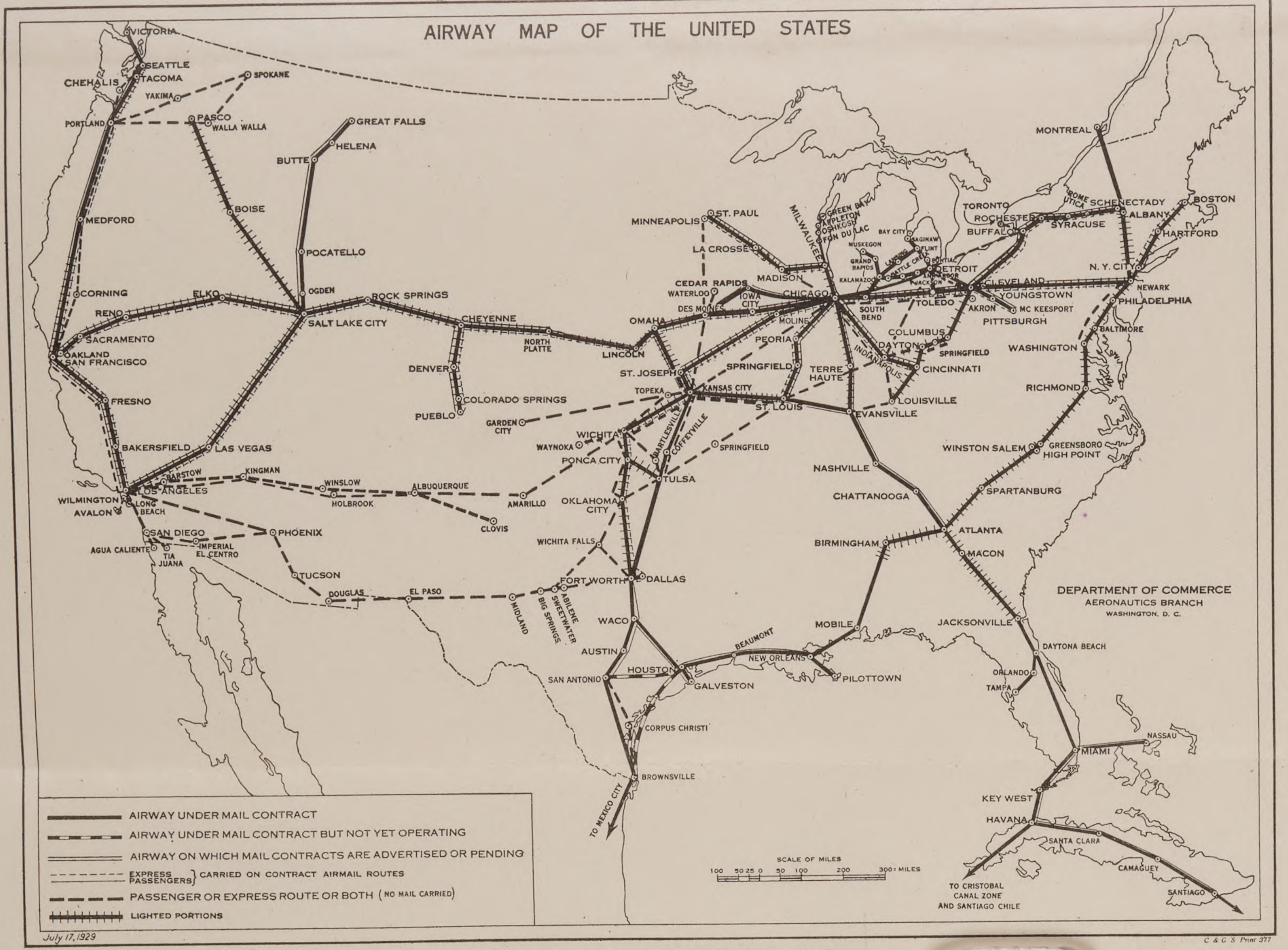
AIRWAY MAP OF THE UNITED STATES
DEPARTMENT OF COMMERCE
AERONAUTICS BRANCH
WASHINGTON, D. C.
AIRWAY UNDER MAIL CONTRACT
AIRWAY UNDER MAIL CONTRACT BUT NOT YET OPERATING
AIRWAY ON WHICH MAIL CONTRACTS ARE ADVERTISED OR PENDING
EXPRESS
PASSENGERS
CARRIED ON CONTRACT AIRMAIL ROUTES
PASSENGER OR EXPRESS ROUTE OR BOTH (NO MAIL CARRIED)
LIGHTED PORTIONS
SCALE OF MILES
100 50 25 0 50 100 200 300 MILES
July 17, 1929
C & G S Print 377
VICTORIA
SEATTLE
TACOMA
CHEHALIS
YAKIMA
SPOKANE
PORTLAND
PASCO
WALLA WALLA
GREAT FALLS
HELENA
BUTTE
BOISE
POCATELLO
OGDEN
MEDFORD
CORNING
ELKO
RENO
SACRAMENTO
OAKLAND
SAN FRANCISCO
SALT LAKE CITY
ROCK SPRINGS
CHEYENNE
NORTH PLATTE
DENVER
COLORADO SPRINGS
PUEBLO
GARDEN CITY
FRESNO
BAKERSFIELD
LAS VEGAS
BARSTOW
KINGMAN
LOS ANGELES
WILMINGTON
LONG BEACH
AVALON
WINSLOW
HOLBROOK
ALBUQUERQUE
AMARILLO
CLOVIS
SAN DIEGO
AGUA CALIENTE
TIA JUANA
IMPERIAL
EL CENTRO
PHOENIX
TUCSON
DOUGLAS
EL PASO
MIDLAND
BIG SPRINGS
SWEETWATER
ABILENE
MINNEAPOLIS
ST. PAUL
LA CROSSE
MADISON
MILWAUKEE
GREEN BAY
APPLETON
OSHKOSH
FON DU LAC
CEDAR RAPIDS
WATERLOO
IOWA CITY
DES MOINES
OMAHA
LINCOLN
ST. JOSEPH
TOPEKA
KANSAS CITY
WICHITA
WAYNOKA
PONCA CITY
BARTLESVILLE
COFFEYVILLE
TULSA
SPRINGFIELD
OKLAHOMA CITY
WICHITA FALLS
FORT WORTH
DALLAS
WACO
AUSTIN
SAN ANTONIO
HOUSTON
GALVESTON
CORPUS CHRISTI
BROWNSVILLE
TO MEXICO CITY
BEAUMONT
NEW ORLEANS
PILOTTOWN
MOBILE
CHICAGO
MOLINE
PEORIA
SPRINGFIELD
ST. LOUIS
TERRE HAUTE
EVANSVILLE
SOUTH BEND
INDIANAPOLIS
LOUISVILLE
CINCINNATI
DAYTON
COLUMBUS
SPRINGFIELD
BAY CITY
SAGINAW
FLINT
MUSKEGON
GRAND RAPIDS
LANSING
BATTLE CREEK
KALAMAZOO
JACKSON
PONTIAC
ANN ARBOR
DETROIT
TOLEDO
CLEVELAND
YOUNGSTOWN
AKRON
MC KEESPORT
PITTSBURGH
TORONTO
ROCHESTER
BUFFALO
SYRACUSE
ROME
UTICA
SCHENECTADY
ALBANY
MONTREAL
BOSTON
HARTFORD
N. Y. CITY
NEWARK
PHILADELPHIA
BALTIMORE
WASHINGTON
RICHMOND
GREENSBORO
HIGH POINT
WINSTON SALEM
SPARTANBURG
NASHVILLE
CHATTANOOGA
ATLANTA
BIRMINGHAM
MACON
JACKSONVILLE
DAYTONA BEACH
ORLANDO
TAMPA
MIAMI
NASSAU
KEY WEST
HAVANA
SANTA CLARA
CAMAGUEY
SANTIAGO
TO CRISTOBAL
CANAL ZONE
AND SANTIAGO CHILE

Fig. 9—Airways map of the United States.

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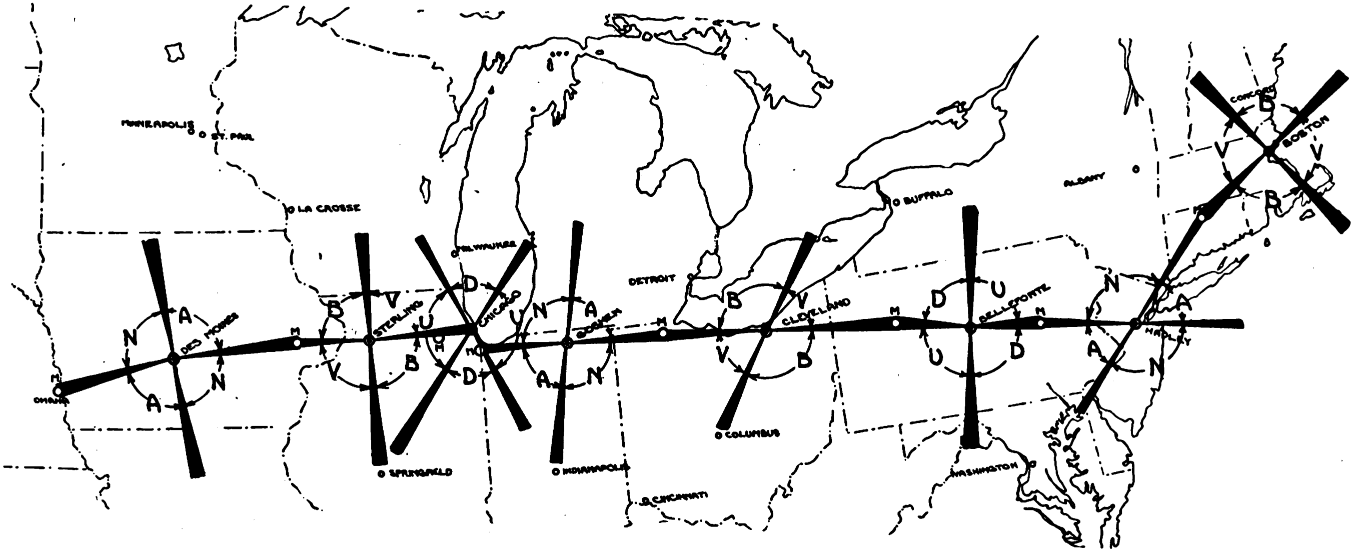
MINNEAPOLIS
ST. PAUL
LA CROSSE
OMAHA
DES MOINES
STERLING
SPRINGFIELD
MILWAUKEE
CHICAGO
GOSHEN
INDIANAPOLIS
DETROIT
CINCINNATI
CLEVELAND
COLUMBUS
BUFFALO
BELLEFONTE
WASHINGTON
HADLEY
ALBANY
BOSTON
CONCORD

Fig. 10—Radio range courses marking a portion of the Transcontinental Airway and important side routes.

will operate a visual indicating instrument on the instrument panel of a plane.[7] A type has been developed to a stage that shows some promise, and arrangements are being made to try this apparatus under actual operating conditions for a sufficient period of time to determine its merit. However, there should be no delay in the installation of directive radio facilities, and the Airways Division is proceeding with the installation of the aural-type of radio range pending the commercial development of the visual type equipment. Equipment for fifteen aural type radio ranges is now under contract for installation in the fall of 1929.

To determine proper operating procedure is probably as important as determining the proper type of equipment. The first method tried was to operate the radio range on a frequency in the radiobeacon band (285–315 kc) and make the telephone broadcasts on a frequency in the 135–350-kc band. This method is not satisfactory since a pilot may be listening to the radio range for directional signals and miss a telephone broadcast containing important information. The method now in use between New York and Chicago is to operate both the radio range and the radio telephone on the same frequency, the radio range operating continuously except during four 3-minute periods at the quarter hours at which time the telephone broadcasts are made. This service is operated continuously throughout the 24 hours.

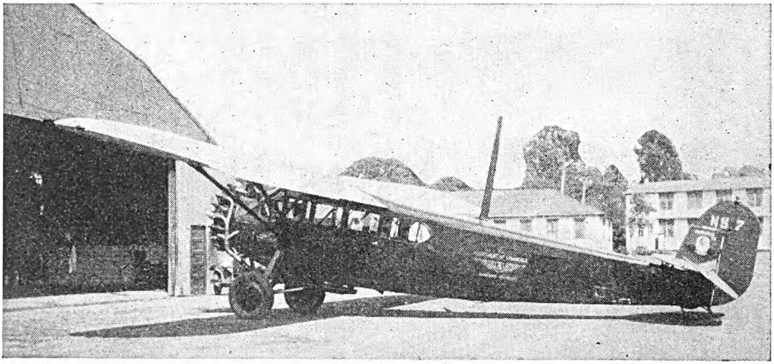

Fig. 11—Airways Division cabin monoplane.

In order that the Airways Division may have full information as to the performance of its various radio and other aeronautical facilities two airplanes have been radio-equipped, one an open cockpit biplane equipped with receiving apparatus only, the second a six-passenger cabin monoplane in which are installed two commercial receiving sets and a 50-watt intermediate-frequency transmitter. This plane is shown in Fig. 11.

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[7] Dellinger and Pratt, "Development of radio aids to air navigation," Proc. I. R. E., 16, 890; July, 1928.

These planes make periodic flights over the civil airways, check up on performance of radio and other facilities, and furnish information regarding the performance of the radio installations particularly. Fig. 12 shows the antenna on the monoplane mentioned above, and is of the vertical type being generally adopted for radio reception. This type of antenna is somewhat ineffective compared to the trailing wire type and reduces the effective range of transmission somewhat. It has the advantage of being permanently in place and reducing certain errors in the reception of radio range signals.

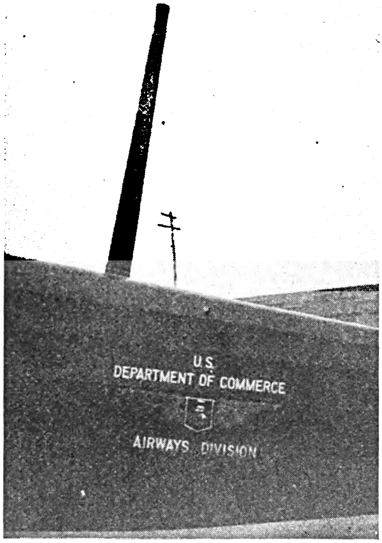

Fig. 12—Receiving antenna on Airways Division monoplane.

The engine ignition system of both planes is shielded with copper braid in the usual manner, and standard type spark plugs are used. The front set of plugs are provided with copper shields to reduce radiation from the plugs themselves. However, experience has shown that when receiving on intermediate frequencies the rear set of plugs may be left unshielded, since they are apparently shielded sufficiently well by the exhaust manifold, cowling, etc.

At the present time there are comparatively few civil airplanes equipped for radio transmission. However, regulations are being considered which will require complete radio installations on certain classes of planes. Briefly the regulations proposed are as follows:

(a) Aircraft carrying one or more passengers in interstate travel shall be equipped with radio receiving apparatus of a satisfactory type so that the benefit of the aeronautical radio facilities on the airway navigated may be secured.

(b) Aircraft carrying six or more passengers in interstate travel shall be provided with radio transmitting and receiving equipment capable of dependable communication with ground stations during the entire period of flight and shall be capable of making a distress call on some designated common frequency assigned as a calling wave. Such plane shall carry a radio operator holding a satisfactory radio license.

These regulations have been submitted to a committee of air transport operators for consideration, and in accordance with established procedure may be put in effect one year after submission.

Consideration is also being given to the inspection and certification of both receiving and transmitting equipment installed on aircraft.

Where air navigation facilities have not been installed by the Department of Commerce, or where present installations do not suffice, it may be necessary to install privately established and maintained radio facilities. The air transport operators have a peculiar responsibility in the carrying of passengers and mail, and the liability for the safety of life and property cannot be transferred to any other agency. The operators have therefore requested the reservation of frequencies for their use through the operation of privately established radio stations. A total of 89 frequency channels have been set up for this service by the Federal Radio Commission, of which 64 are in the 1500–6000-kc band. These frequencies are being assigned to the various air routes. That is, a particular channel is used along a given route regardless of the number of transport companies flying that route. All companies have equal rights now and in the future as to this service and provide their own installations jointly, sharing equally in the installation costs, operation costs, and liability.

The following important distress and calling frequencies have been established.

278 kc	Calling and working frequency used by airports only. Power not to exceed 10 watts.
333 kc	International air calling frequency.

375 kc	Radiocompass service (mobile transmitters, fixed compass on the ground).
500 kc	International calling and distress frequency for ships and aircraft over the seas.
3106 kc	National calling frequency for all transport and itinerant airplanes.

Since communications to airplanes under conditions of poor visibility must necessarily be made by radio, it will be necessary to provide for communicating landing directions at airports by means of the radiotelephone. The frequency of 278 kc mentioned above was set up by the Federal Radio Commission for this purpose. The radio transmitters at airports must have limited power and a range not in excess of five miles so as to prevent interference between neighboring airports. Under this plan of having all airport transmitters on the same frequency, the pilot will listen in for landing directions upon approaching an airport and will follow the orders as to landing procedure received from the airport manager.

There can be no doubt that air transportation from now on will depend in a large measure upon radio for its safety and flying efficiency. The radio facilities that are now being established have been planned so that all classes of fliers may make use of the service by installing a simple receiving set on the airplane, and receive practically all of the information required to fly the civil airways safely. The comprehensive weather service that is now being established, and the facilities provided to make this information immediately available to the pilot in the air, are important. In addition to the weather broadcasting service, and probably of equal importance, are the directive radio signals. These will indicate to the pilot his exact course, and should he deviate, the signals will permit him to locate the course again and bring his flight to a successful conclusion at the airport. Accidents due to weather conditions as well as other causes will thus be greatly reduced.

* * *

Proceedings of the Institute of Radio Engineers
Volume 17, Number 12 *December, 1929*

APPLYING THE VISUAL DOUBLE-MODULATION TYPE RADIO RANGE TO THE AIRWAYS*

By
H. Diamond
(Bureau of Standards, Washington, D. C.)

Summary—*This paper deals with methods for aligning the courses of the visual radio range with the fixed airways. It has previously been shown that the courses of the aural radio range may be shifted by the use of a vertical wire antenna in conjunction with the transmitting loop antennas or by varying the relative power in the two antennas. These methods are, in part, applicable to the visual system. In the aural system the goniometer primaries are excited alternately. This permits independent consideration of the field patterns due to the primaries. In the visual system this is not the case as both goniometer primaries are excited all the time. Two cases present themselves, the condition when the currents in the primaries are in time phase and the condition when they are in quadrature time phase. The former condition results in two beacon courses which are 180 deg. apart and cannot be shifted from this relationship. The latter condition yields four beacon courses. A mathematical analysis is made of this case and the amounts of angular variation possible using several methods of attack are tabulated.*

A method of obtaining small amounts of shift by an adjustment of the receiving equipment aboard the airplanes is also described; one of the reeds is shunted by a suitable resistance in order that the reeds will vibrate equally when on one side of the equisignal zone. This method permits a great flexibility in securing a desired course and is suitable only for employment with the visual system. Sample calculations are made for actual airway routes to demonstrate the several methods of attack.

I. Introduction

Methods for aligning the courses of the aural type radio range (i.e., directive radiobeacon) with the fixed airways are described in a paper by Kear and Jackson.[1] Extensions of these methods may be used to make the visual double-modulation type radio range equally flexible. This paper describes a number of circuit arrangements, the application of which makes possible the use of a single visual type radio range for serving two, three, or four courses radiating from a given airport at arbitrary angles with each other.

The procedure of aligning the courses of the visual type radio range with the fixed airways is necessarily somewhat different from that followed in the aural type, owing to the essential difference between the signals used for marking out the beacon courses in the two systems. With the aural system, a pilot's position on either side of the course is indicated by the relative strength of two Morse characters. When on the course these characters blend into one long dash of constant signal strength. In order to obtain proper interlocking no portions of the two characters used can be transmitted simultaneously. With the visual system, a course is indicated by the equality of vibration amplitude of two reeds mechanically tuned to the two modulation frequencies used at the beacon. For maximum reed amplitude, it is desirable to provide continuous modulation at both modulation frequencies. The two distinguishing signals of the beacon are, therefore, always sent out simultaneously, resulting, under normal conditions, in a space pattern quite different from that obtained from an aural radio range.[2] Two, rather than four, beacon courses are obtained. To produce four beacon courses it is necessary to advance the time phase of one of the carrier frequencies of the system 90 deg. beyond that of

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* Dewey decimal classification: R526.1. Original manuscript received by the Institute, October 8, 1929. Published by permission of the Director of the Bureau of Standards of the U. S. Department of Commerce. To be published in a forthcoming issue of the Bureau of Standards Journal of Research.

[1] Kear and Jackson, "Applying the radio range to the airways," this issue of the Proc., page 2268. Also to be published in the Bureau of Standards Journal of Research.

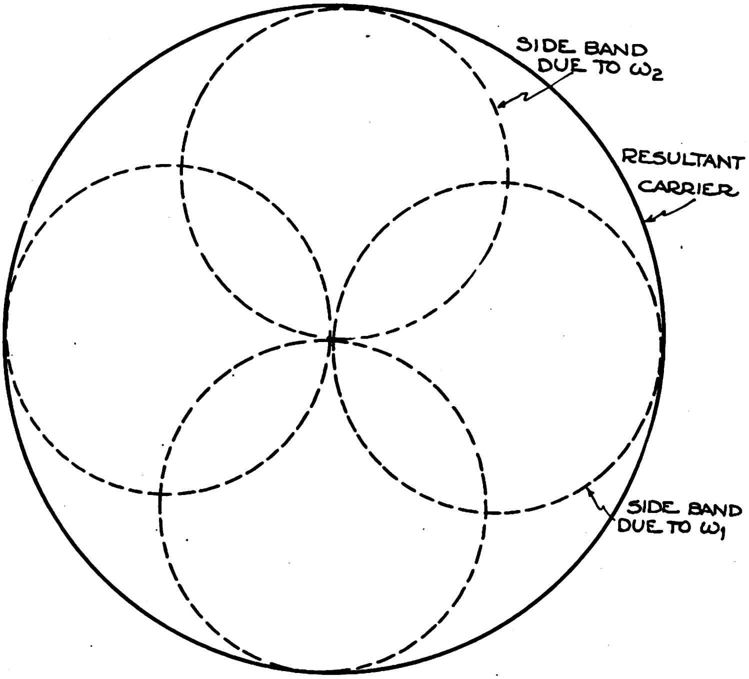

Fig. 1—Space pattern radiated by double-modulation type radio range when antenna currents are 90 deg. out of phase.

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[2] This was shown by Pratt in Fig. 2 of his paper on "Field intensity characteristics of double modulation type of direction radio beacon," Proc. I. R. E., 17, 875; May, 1929.

the other. The space pattern is then as indicated in Fig. 4, page 877, of Pratt's paper, the four beacon courses being displaced by 90 deg. from each other. For convenience in reference, this figure is reproduced here as Fig. 1. The received polar pattern is shown in Fig. 2. The trigonometric expression for the beacon space pattern is given in equation (1).[3] The corresponding expression for the polar pattern as received on the reeds is given in equation (2), which assumes square-law detection.

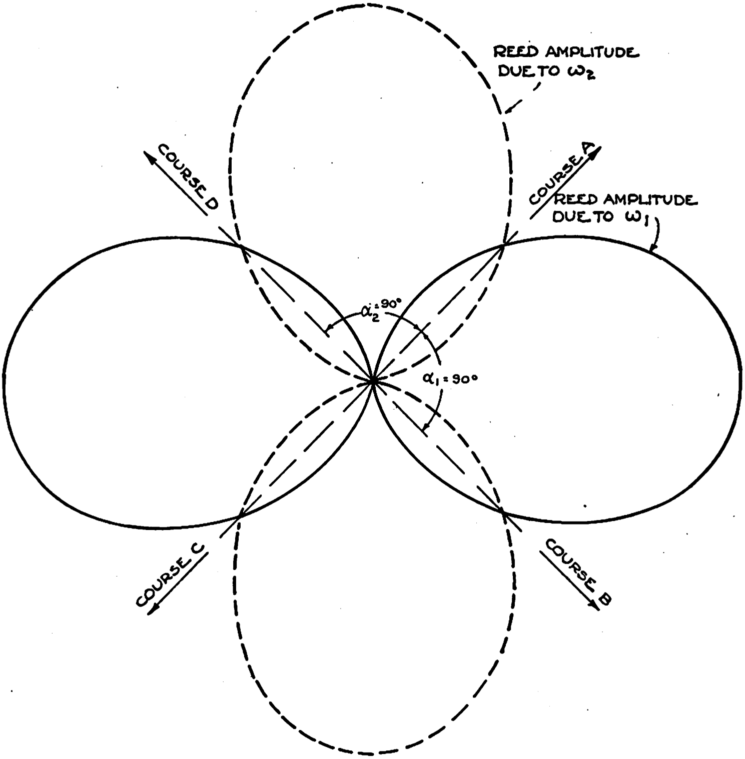
REED AMPLITUDE DUE TO ω_2
COURSE D
COURSE A
REED AMPLITUDE DUE TO ω_1
$\alpha_2=90°$
$\alpha_1=90°$
COURSE C
COURSE B

Fig. 2—Received polar pattern corresponding to the space pattern of Fig. 1.

$$e_p = K \left\{ \begin{array}{l} E_0[\cos \omega t \sin \theta + \sin \omega t \cos \theta] \\ -\dfrac{E_1}{2}[\sin (\omega-\omega_1)t - \sin (\omega+\omega_1)t] \sin \theta \\ +\dfrac{E_2}{2}[\cos (\omega-\omega_2)t - \cos (\omega+\omega_2)t] \cos \theta \end{array} \right\} \quad (1)$$

[3] This is equivalent to equation (5), page 876 of Pratt's paper.

where e_p is the field intensity at any point P in space as a polar function of the angle θ, and, as is apparent, consists of a carrier and two sets of sidebands; $E_1/E_0 \times 100$ is the percentage modulation in amplifier branch 1 due to ω_1; $E_2/E_0 \times 100$ is the percentage modulation in amplifier branch 2 due to ω_2. In this and all the following equations we will assume $E_1 = E_2 = E_0$ unless otherwise stated.

$$e_r = KK'E_0\left\{E_1 \sin \omega_1 t \sin^2 \theta + E_2 \sin \omega_2 t \cos^2 \theta\right\} \tag{2}$$

where e_r is the received signal strength at any point P as a polar function of the angle θ. A course occurs whenever the two terms to the right of the equality sign of equation (2) are equal.

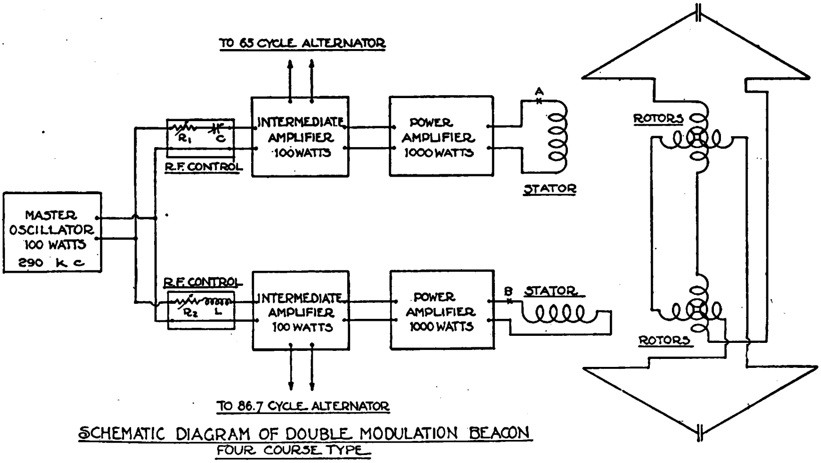
TO 65 CYCLE ALTERNATOR
R1
C
R.F. CONTROL
INTERMEDIATE AMPLIFIER 100 WATTS
POWER AMPLIFIER 1000 WATTS
A
STATOR
ROTORS
MASTER OSCILLATOR 100 WATTS 290 K C
R.F. CONTROL
R2
L
INTERMEDIATE AMPLIFIER 100 WATTS
POWER AMPLIFIER 1000 WATTS
B
STATOR
ROTORS
TO 86.7 CYCLE ALTERNATOR
SCHEMATIC DIAGRAM OF DOUBLE MODULATION BEACON FOUR COURSE TYPE

Fig. 3—Schematic diagram of double-modulation radio range—four-course type.

From a study of equations (1) and (2), it is interesting to note that the useful signal due to ω_1 is the result of the beating of the carrier of branch 1 (*not* the resultant carrier) with the sidebands transmitted by branch 1. Similarly, the useful signal due to ω_2 is the result of the beating of the carrier and sidebands transmitted by branch 2. These relationships are true, however, only in the special case when the time phase between the two carriers of the system is equal to 90 deg.

A schematic diagram of the circuit arrangement used for obtaining four beacon courses at 90 deg. with each other is shown in Fig. 3. The means for producing a 90-deg. time-phase displacement between the two carrier-frequency currents of the system consists simply of inserting a suitable capacitive reactance in the supply lead from the master oscillator to one amplifier branch and a suitable inductive reactance in the supply lead to the other amplifier branch. For convenience in phase adjustment the condenser is made variable, and a variable resistor is connected in series with each reactance.

II. METHOD A. SHIFTING COURSES BY AMPLITUDE REDUCTION

One method of shifting the four beacon courses from their 90-deg. space relationship is to reduce the percentage modulation of one amplifier branch, keeping the magnitude of the carrier unchanged. The space pattern radiated by the beacon is then as shown in Fig. 4, and the received polar diagram as shown in Fig. 5. A 30 per cent reduction in the percentage modulation of one amplifier branch is assumed. The trigonometric equations for the radiated space pattern and received polar pattern are then, respectively:

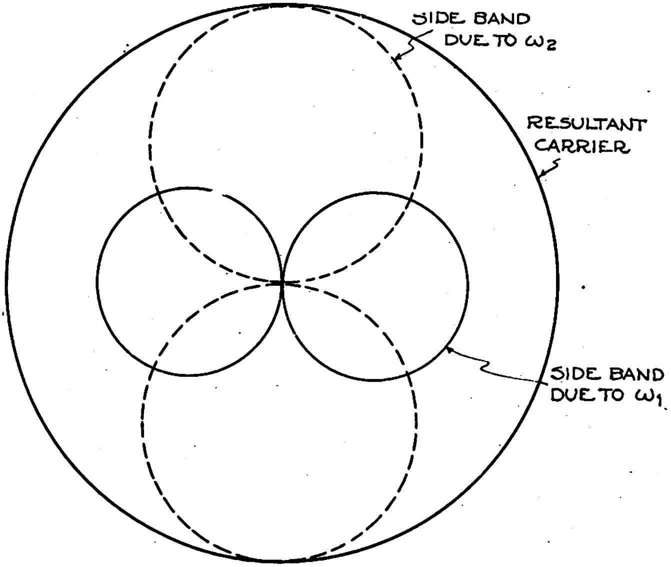

Fig. 4—Space pattern when percentage modulation in one amplifier branch is reduced.

$$\begin{aligned} e_p = K\Big[E_0\{\cos\omega t\sin\theta+\sin\omega t\cos\theta\} \\ -\frac{0.7\,E_1}{2}\{[\sin(\omega-\omega_1)t-\sin(\omega+\omega_1)t]\sin\theta\} \\ +\frac{E_2}{2}\{[\cos(\omega-\omega_2)t-\cos(\omega+\omega_2)t]\cos\theta\}\Big] \end{aligned} \tag{3}$$

$$e_r = KK'E_0[0.7\,E_1\sin\omega_1 t\sin^2\theta+E_2\sin\omega_2 t\cos^2\theta] \tag{4}$$

A course occurs whenever the two terms to the right of the equality sign of equation (4) are equal. It will be observed from Fig. 5 that the two sets of 180-deg. courses, (A, B) and (C, D), are now displaced by $\alpha_1 = 80$ deg. and $\alpha_2 = 100$ deg. as compared with $\alpha_1 = \alpha_2 = 90$ deg. when the degree of modulation in the two amplifier branches is the same.

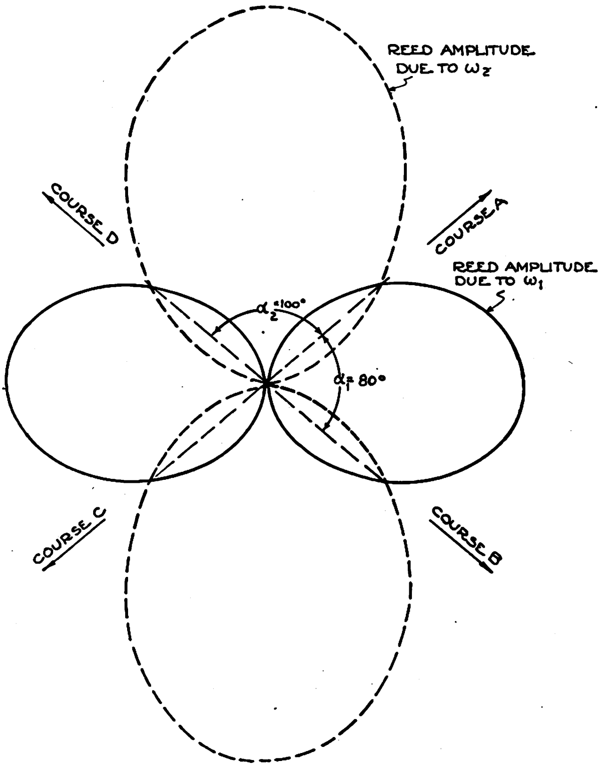

Fig. 5—Received polar pattern corresponding to Fig. 4.

The method of modulation adopted in the visual radio range transmitter, consisting of plate excitation of the intermediate amplifiers from sources of a-c supply of suitable frequencies, does not permit an easy adjustment of the percentage modulation. It is more convenient to reduce the magnitude of both carrier and sidebands of one amplifier branch. This is accomplished by increasing the value of the resistor R_1 of the radio-frequency control (Fig. 3) and readjusting C to keep the phase of the voltage applied to the grid of the intermediate amplifier tube unchanged. Assuming a 30 per cent reduction in the amplitude of carrier and sidebands of one amplifier branch, the radiated space pattern becomes that of Fig. 6, while the polar diagram as received on the reeds is shown in Fig. 7. The trigonometric expression for the space pattern is given in equation (5).

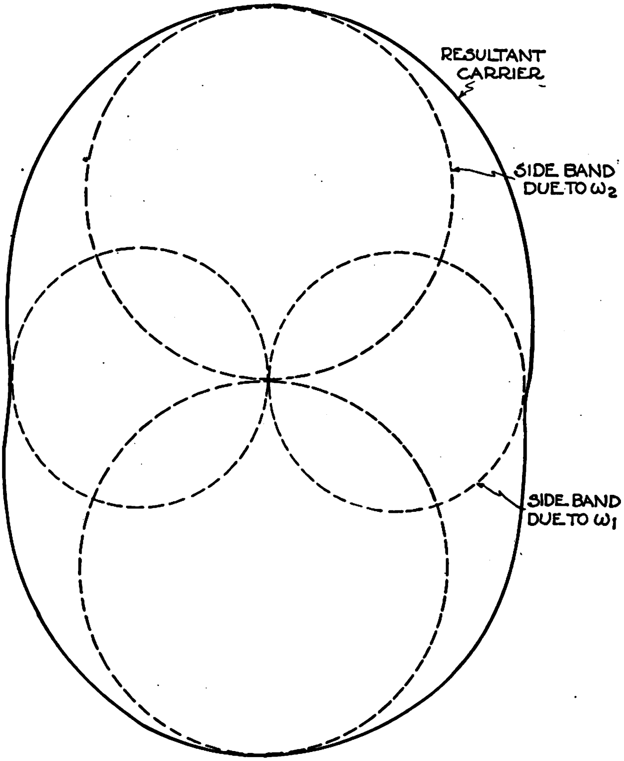

Fig. 6—Space pattern when magnitude of the carrier and sidebands transmitted by one amplifier branch is reduced.

$$\begin{aligned} e_p = K\Big[E_0\{0.7 \cos \omega t \sin \theta + \sin \omega t \cos \theta\} \\ -0.7\frac{E_1}{2}\{[\sin (\omega-\omega_1)t - \sin (\omega+\omega_1)t] \sin \theta\} \\ +\frac{E_2}{2}\{[\cos (\omega-\omega_2)t - \cos (\omega+\omega_2)t] \cos \theta\}\Big]. \end{aligned} \tag{5}$$

The corresponding expression for the received signal becomes

$$e_r = KK'E_0[0.49E_1 \sin \omega_1 t \sin^2 \theta + E_2 \sin \omega_2 t \cos^2 \theta]. \tag{6}$$

A course will occur whenever the two reed deflections are of equal magnitude, i.e., when

$$\left|0.7\sin\theta\right| = \left|\cos\theta\right|. \tag{7}$$

Equation (7) follows from the facts that the observed reed deflections are determined by the maximum instantaneous values of $E_1 \sin \omega_1 t$ and $E_2 \sin \omega_2 t$ and that the reeds are adjusted for equal sensitivity. Note that the courses have been shifted in greater amount than in Fig. 5. The values for the angles α_1 and α_2 are now 70 deg. and 110 deg., respectively. The relative reduction in strength of signal received when on course is, however, also greater.

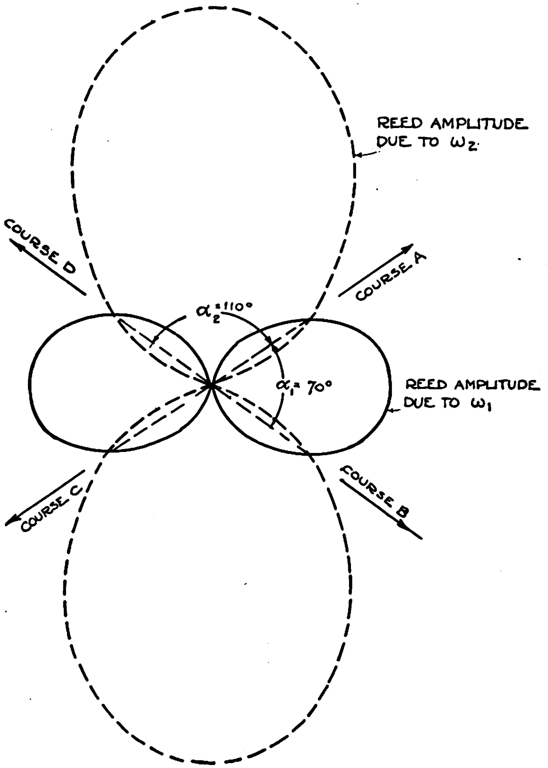
REED AMPLITUDE DUE TO ω_2
COURSE D
COURSE A
$\alpha_2 = 110°$
$\alpha_1 = 70°$
REED AMPLITUDE DUE TO ω_1
COURSE C
COURSE B

Fig. 7—Received polar pattern corresponding to Fig. 6.

Two factors determine the minimum value for α_1 (or the maximum value for α_2) that can be obtained: (a) the signal strength received when on any given course should not be reduced below 50 per cent of that received when the four courses are displaced by 90 deg.; (b) a deviation of at least 20 deg. on either side of the course should be possible without losing indications as to the direction back to the course. Keeping in mind these requirements, both methods described above yield as tolerable limits ($\alpha_1 = 60$ deg., $\alpha_2 = 120$ deg.). Using these methods, then, it becomes possible to fit two of the beacon courses to any two airways separated by an angle in the range 60 deg. to 120 deg.

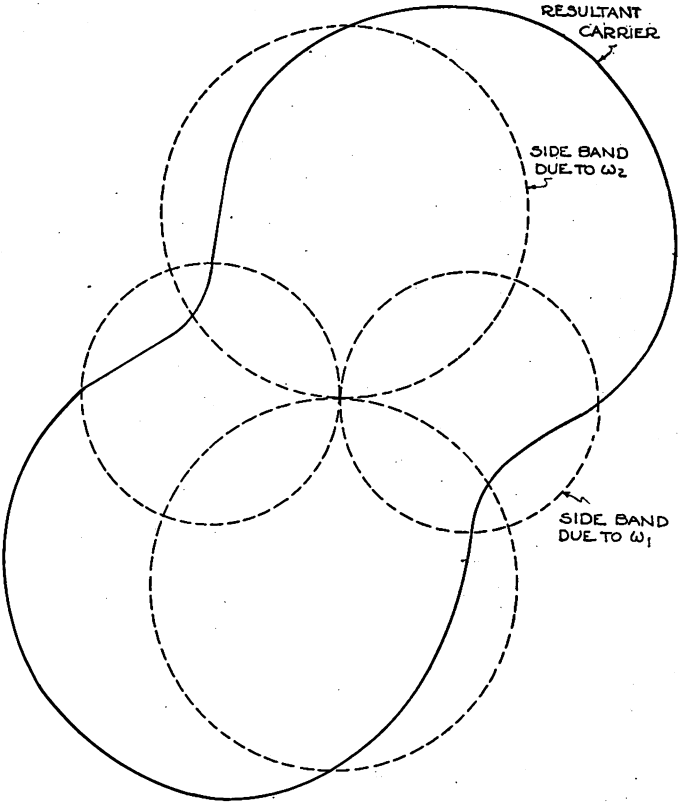

Fig. 8—Space pattern for same conditions as in Fig. 6 except that antenna currents are 45 deg. out of time phase instead of 90 deg.

Before passing on to further methods for making the other two beacon courses useful, it is of interest to study the effect of changing the time phase displacement between the two modulated waves of the transmitting system. As noted above, the difference in time phase is normally adjusted to 90 deg. Suppose that the time phase difference is made 45 deg. (by adjusting L and C, Fig. 3). For the condition that the magnitude of carrier and sidebands transmitted by one amplifier branch is reduced to 70 per cent that of the other amplifier branch, the beacon space pattern becomes that shown in Fig. 8 and the received polar pattern as shown in Fig. 9. The expression for the radiated space pattern under these conditions is given in (8) and for the received polar pattern in (9).

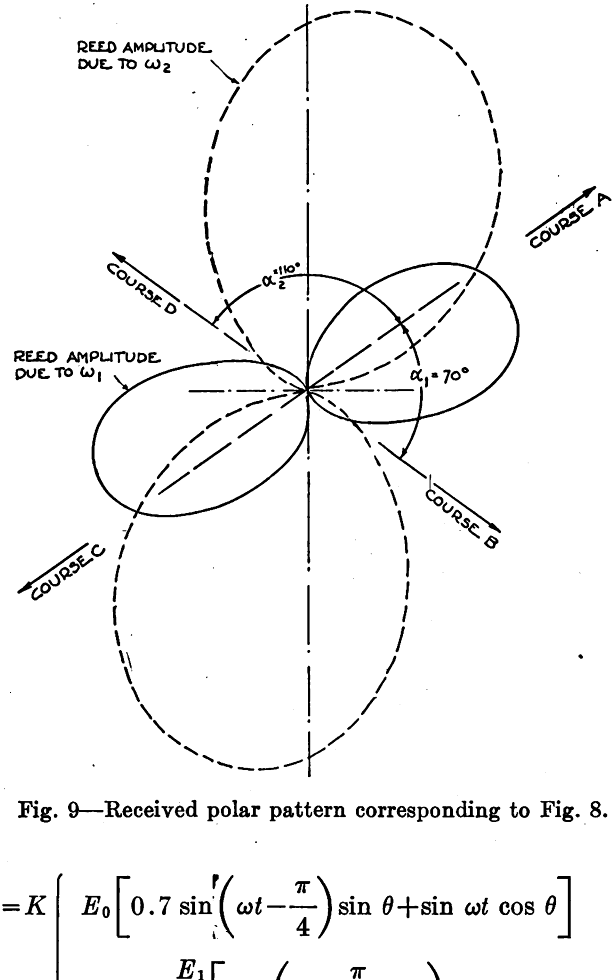
REED AMPLITUDE DUE TO ω_2
COURSE A
COURSE D
$\alpha_2 = 110°$
REED AMPLITUDE DUE TO ω_1
$\alpha_1 = 70°$
COURSE B
COURSE C

Fig. 9—Received polar pattern corresponding to Fig. 8.

$$e_p = K\left\{\begin{aligned} & E_0\left[0.7\sin\left(\omega t-\frac{\pi}{4}\right)\sin\theta+\sin\omega t\cos\theta\right] \\ & +0.7\frac{E_1}{2}\left[\cos\left(\omega t-\frac{\pi}{4}-\omega_1 t\right)\right. \\ & \qquad\qquad \left.-\cos\left(\omega t-\frac{\pi}{4}+\omega_1 t\right)\right]\sin\theta \\ & +\frac{E_2}{2}\left[\cos(\omega-\omega_2)t-\cos(\omega+\omega_2)t\right]\cos\theta \end{aligned}\right\} \tag{8}$$

$$e_r = KK'E_0\left[E_1 \sin \omega_1 t \left\{0.7 \times \frac{1}{\sqrt{2}} \sin\theta \cos\theta + 0.49 \sin^2\theta\right\} + E_2 \sin \omega_2 t \left\{0.7 \times \frac{1}{\sqrt{2}} \sin\theta \cos\theta + \cos^2\theta\right\}\right] \tag{9}$$

Note that since the two carrier frequency currents of the system are no longer displaced in time phase by 90 deg., the relationship observed in connection with equations (1) and (2) no longer holds.

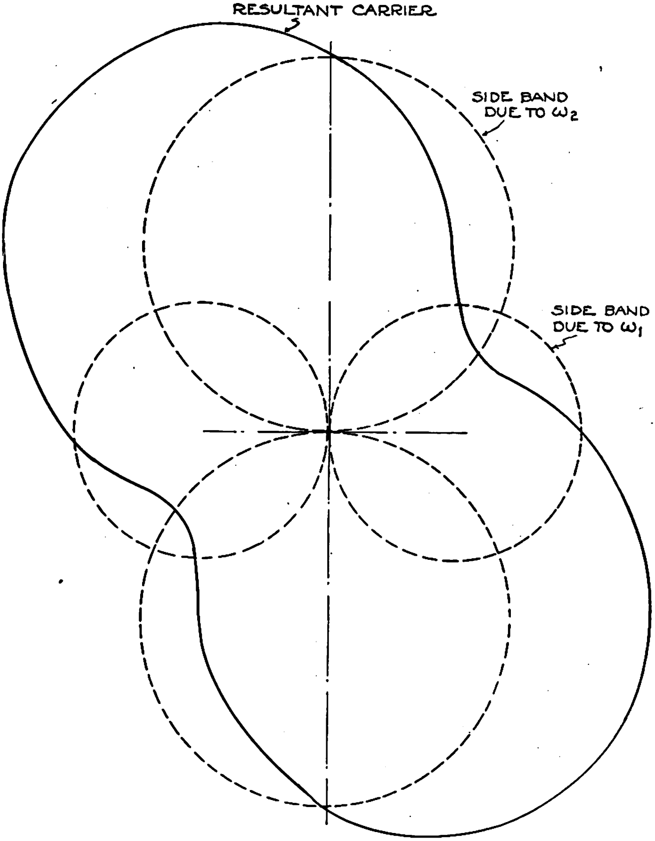

Fig. 10—Space pattern for same conditions as in Fig. 6 except that antenna currents are 135 deg. out of time phase instead of 90 deg.

If the time phase displacement is made 135 deg., the resultant space pattern is as shown in Fig. 10, and the received polar pattern as shown in Fig. 11. The trigonometric equations corresponding to these patterns are similar to (8) and (9) and will not be given here. Note that a change in the time phase displacement does not result in a shifting of the beacon courses. An exact adjustment of time phase displacement to 90 deg. is therefore not necessary. By varying this displacement, however, it is possible to control at will the relative signal strength on the two sets of 180-deg. courses. In this way, if the two airways to be served by a given radio range differ in length, a stronger course signal may be directed along the longer airway.

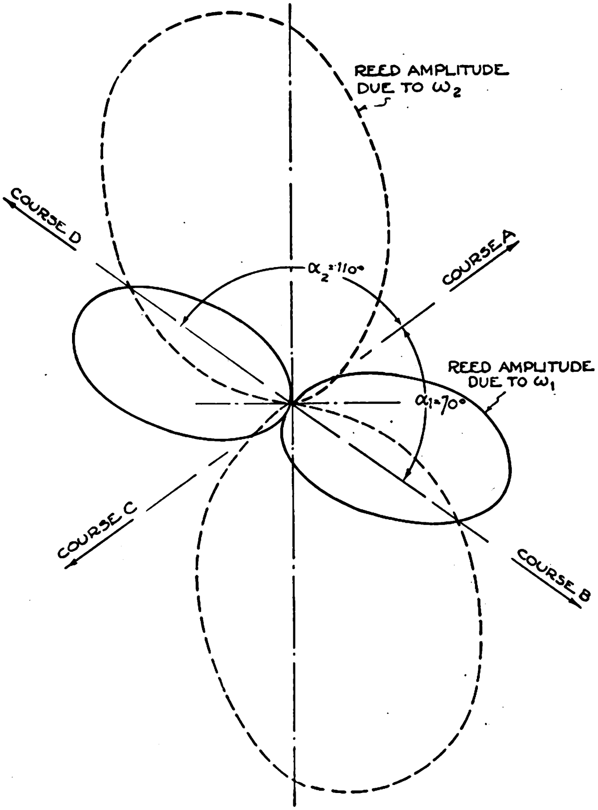
REED AMPLITUDE DUE TO ω_2
COURSE D
COURSE A
$\alpha_2=110°$
REED AMPLITUDE DUE TO ω_1
$\alpha_1=70°$
COURSE C
COURSE B

Fig. 11—Received polar pattern corresponding to Fig. 10.

III. METHOD B. USE OF AUXILIARY VERTICAL ANTENNA FOR SHIFTING COURSES

A further method for shifting the courses from their 90-deg. space relationship consists of supplying a circular radiation of the carrier and sidebands transmitted by one amplifier branch, in addition to the figure-of-eight radiation due to the loop antenna fed by that amplifier branch. A vertical antenna, running the length of the beacon tower and inductively coupled to the output circuit of one of the two amplifying branches of the transmitting system, is employed for obtaining this additional radiation. Assuming a ratio of amplitude of circular radiation to maximum amplitude of figure-of-eight radiation equal to 0.28, the radiated space pattern becomes as shown in Fig. 12, and the received polar diagram as shown in Fig. 13. The expression for the radiated space pattern is given in (10) and for the received pattern in (11).

$$e_p = K\left\{\begin{array}{l} E_0[\cos \omega t \sin \theta + \sin \omega t(0.28 + \cos \theta)] \\ -\dfrac{E_1}{2}[\sin (\omega - \omega_1)t - \sin (\omega + \omega_1)t] \sin \theta \\ +\dfrac{E_2}{2}[\cos (\omega - \omega_2)t - \cos (\omega + \omega_2)t]\,[0.28 + \cos \theta] \end{array}\right\} \tag{10}$$

$$e_r = KK'E_0\{E_1 \sin \omega_1 t \sin^2 \theta + E_2 \sin \omega_2 t[0.28 + \cos \theta]^2\} \tag{11}$$

A beacon course will occur when (12) is satisfied.

$$|\sin \theta| = |(0.28 + \cos \theta)|. \tag{12}$$

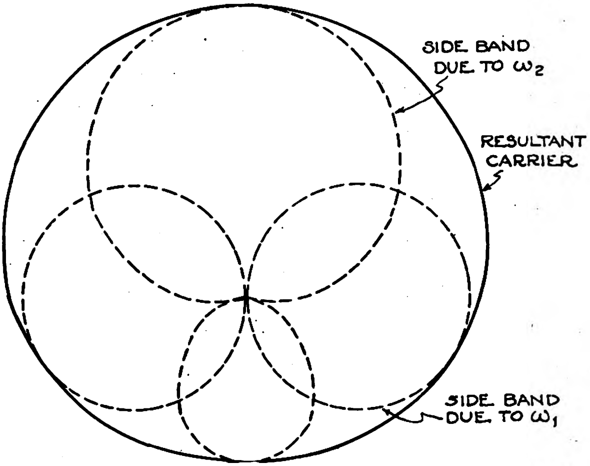

Fig. 12—Space pattern when circular radiation of the modulated wave transmitted by one amplifier branch is added to the normal figure-of-eight radiation due to that branch.

In (10) the current in the vertical antenna is taken as in time phase with the current in loop antenna (2). This is the condition of maximum course shift for a given vertical antenna current. If these two currents were in time quadrature, no course shift would occur. It is, therefore, important to insure that these currents are in phase;

(a) by keeping the vertical antenna and loop antenna (2) accurately in tune; and (b) by connecting the primary of the coupling transformer to the vertical antenna in the output circuit of the same amplifier stage in which the primary winding of the coupling arrangement to loop antenna (2) is connected (i.e., at point *B*, Fig. 3).

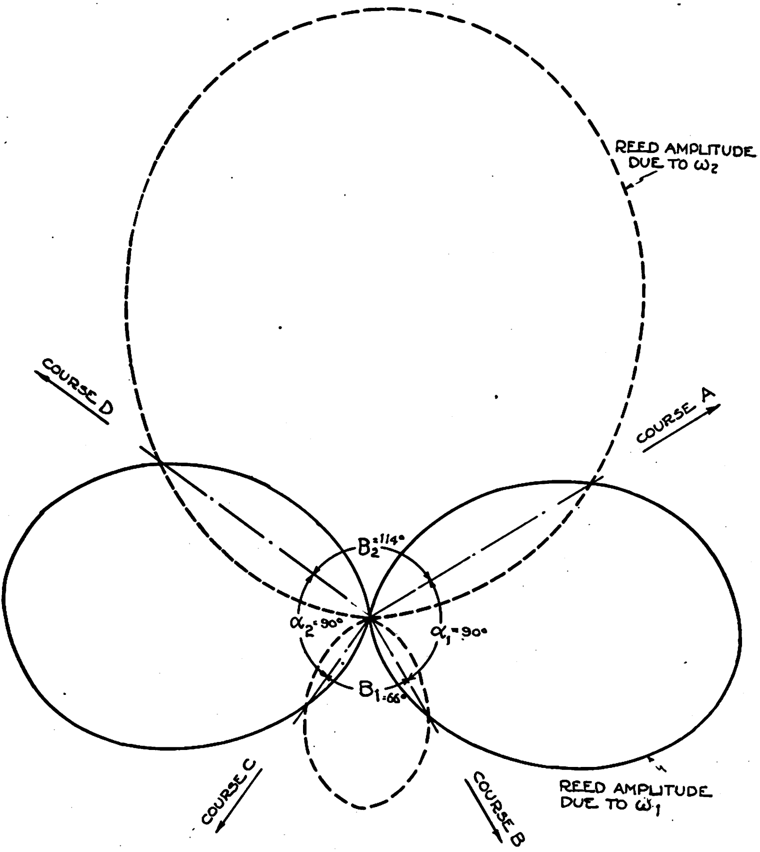

Fig. 13—Received polar pattern corresponding to Fig. 12.

Comparing the received polar diagram of Fig. 13 with that for the beacon normally adjusted (see Fig. 2), it will be observed that the addition of a circular component to the normal radiation characteristic of the modulated wave due to ω_2 results in decreasing the angle β_1 between courses *B* and *C* and increasing the angle β_2 between courses *D* and *A*. The angles α_1 between courses *A* and *B* and α_2 between courses *C* and *D* remain equal to 90 deg. Applying the same criterions as were used in the method of course shifting by amplitude reduction, (namely, that the signal strength on course should not be reduced below 50 per cent of normal, and that a deviation of 20 deg. on either side of a course be possible), the minimum tolerable value for β_1 is 60 deg. and the maximum value for β_2 is 120 deg.

A comparison of methods A and B for varying the angle between two beacon courses is of interest. Using method A, two airways separated by any angle in the range 60 deg. to 120 deg. may be served. This is also true using method B. In method A, however, each of the two remaining beacon courses is displaced by 180 deg. from one of the two courses used, while in method B each of the two remaining courses is displaced by 90 deg. from one of the two courses employed.

IV. METHOD C. COMBINATION OF METHODS A AND B

Methods A and B may be combined, yielding an arrangement for serving three airways simultaneously, provided the angles between these airways are within the limits tabulated below. For convenience in tabulation, the three airways are called A, B, and C', respectively. Any two of these (say A and B) may be from 60 deg. to 90 deg. apart. The third airway (C') may then be disposed from either A or B by any angle within the range given in the table. The two criterions noted above under methods A and B are fulfilled in this table. As shown, the practicable range of angles between C' and either A or B depends upon the angle between A and B. A greater variation can of course be obtained if it is permissible to reduce the signal strength received when on course. The third column of Table I shows the possible range if the minimum permissible signal strength when on any given course is taken as 33 per cent of normal.

TABLE I

Angle between A and B	Range of Permissible Angles between C' and either A or B if $E_r'/E_r=0.5$	Range of Permissible Angles between C' and either A or B if $E_r'/E_r=0.33$
60 deg.	120 deg., 180 deg.	90 deg.–210 deg.
65 deg.	105 deg.–125 deg., 170 deg.–190 deg.	80 deg.–215 deg.
70 deg.	90 deg.–130 deg., 160 deg.–200 deg.	70 deg.–220 deg.
75 deg.	82.5 deg.–127.5 deg., 157.5 deg.–202.5 deg.	65 deg.–220 deg.
80 deg.	75 deg.–125 deg., 155 deg.–205 deg.	60 deg.–220 deg.
85 deg.	67.5 deg.–122.5 deg., 152.5 deg.–207.5 deg.	55 deg.–135 deg., 140 deg.–220 deg.
90 deg.	60 deg.–120 deg., 150 deg.–210 deg.	50 deg.–130 deg., 140 deg.–220 deg.

The results indicated in Table I are obtained by reducing the amplitude of carrier and sidebands transmitted by one amplifier branch to obtain any angle from 60 deg. to 90 deg. between courses A and B, and then introducing a circular radiation of the carrier and sidebands transmitted by the other amplifier branch in addition to the normal figure-of-eight radiation for that branch. The space pattern as radiated from the beacon then becomes as shown in Fig. 14, and the received polar diagram in Fig. 15. The trigonometric equation for the radiated space pattern is given by (13), and for the received pattern by (14).

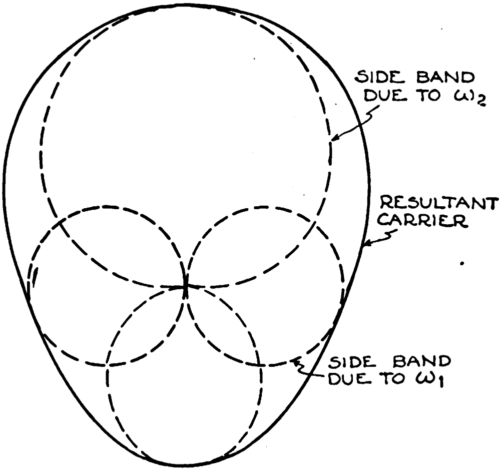

Fig. 14—Space pattern when methods *A* and *B* for course-shifting are combined.

$$e_p = K\left\{\begin{array}{l} E_0[C_1 \cos \omega t \sin \theta + \sin \omega t(K_2 + \cos \theta)] \\ -C_1\dfrac{E_1}{2}[\sin(\omega-\omega_1)t - \sin(\omega+\omega_1)t]\sin\theta \\ +\dfrac{E_2}{2}[\cos(\omega-\omega_2)t - \cos(\omega+\omega_2)t][K_2+\cos\theta] \end{array}\right\} \qquad (13)$$

where $K_2 = \dfrac{\text{amplitude of circular radiation due to } \omega_2}{\text{maximum amplitude of figure-of-eight radiation due to } \omega_2}$

C_1 = reduction factor for carrier and sidebands due to ω_1

$$e_r = KK'E_0\{C_1^2E_1 \sin \omega_1 t \sin^2\theta + E_2 \sin \omega_2 t[K_2 + \cos \theta]^2\} \qquad (14)$$

In Figs. 14 and 15, $C_1 = 0.7$ and $K_2 = 0.22$.

Referring to (14), a course will occur whenever

$$|K_2 + \cos \theta| = |C_1 \sin \theta|. \qquad (15)$$

This equation will have four solutions, one for each quadrant. In the first quadrant

$$K_2 + \cos \theta_1 = C_1 \sin \theta_1 \qquad \theta_1 = \text{angle of course } A \qquad (16)$$

In the second quadrant

$$K_2 - \cos \theta_2 = C_1 \sin \theta_2 \qquad \theta_2 = \text{angle of course } B \qquad (17)$$

In the third quadrant

$$K_2 - \cos\theta_3 = -C_1 \sin\theta_3 \qquad \theta_3 = \text{angle of course } C \tag{18}$$

In the fourth quadrant

$$K_2 + \cos\theta_4 = -C_1 \sin\theta_4 \qquad \theta_4 = \text{angle of course } D \tag{19}$$

In (16), (17), (18), and (19) the factors $\sin\theta_1$, $\cos\theta_1$, etc., are constants of positive sign.

Subtracting (17) from (16) and solving, we obtain

$$\theta_2 - \theta_1 = 2\alpha \qquad \text{where } \alpha = \tan^{-1} C_1 \tag{20}$$

Similarly, subtracting (19) from (18) and solving, we obtain

$$\theta_4 - \theta_3 = 2\alpha. \tag{21}$$

We note, then, that the angle between courses A and B, and likewise between courses C and D, are dependent upon C_1, the reduction factor of carrier and sidebands in one amplifier branch. These angles are independent of the amount of circular radiation of carrier and sidebands added to the normal figure-of-eight radiation of the other amplifier branch.

Again, subtracting (18) from (17), and solving, we obtain

$$\theta_3 = -\theta_2. \tag{22}$$

Similarly, subtracting (16) from (19) and solving, we obtain

$$\theta_4 = -\theta_1. \tag{23}$$

Equations (22) and (23) indicate that courses B and C, and also courses D and A, are symmetrical about the 0-deg.–180-deg. axis. This may be seen by reference to Fig. 15. The angle between courses B and C is decreased, and the angle between courses D and A increased in like amount, as the ratio of maximum amplitude of circular to figure-of-eight radiation due to the second amplifier branch is increased. If the phase of the circular radiation is reversed with respect to the phase of the figure-of-eight radiation, an increase in this ratio will result in an increase in the angle between courses B and C and a decrease in the angle between D and A. In either case, for a given fixed value of C_1, the angles between the courses A and B and between C and D remain fixed at a value $2\tan^{-1} C_1$.

The procedure of determining the proper values for C_1 and for K_2 in order to serve three courses at given angles with each other is as follows:

(a) Suppose that courses A and B are 75 deg. apart.
Place $2\tan^{-1} C_1 = 75$ deg.
Then $C_1 = \tan 37.5 \text{ deg.} = 0.767$.

(b) Now, suppose that course C' is 90 deg. from course B. For convenience in fixing ideas, refer to Fig. 15. Under normal adjustments, i.e., with no additional vertical radiation, course C would be 105 deg. from course B. To decrease this angle to 90 deg., circular radiation of proper phase must be introduced. Since B and C are symmetrically disposed about the 0-deg.–180-deg. axis, and the angle between courses B and C equals 90 deg., the angle of course C must be 225 deg. Place

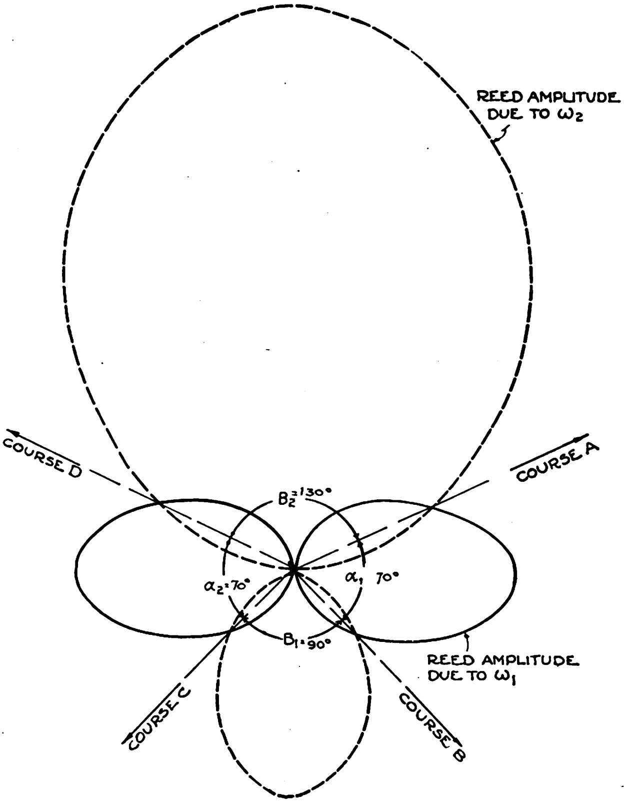
REED AMPLITUDE DUE TO ω_2
COURSE D
COURSE A
$B_2 = 130°$
$\alpha_2 = 70°$
α_1 70°
$B_1 = 90°$
REED AMPLITUDE DUE TO ω_1
COURSE C
COURSE B

Fig. 15—Received polar pattern corresponding to Fig. 14.

$$\left| K_2 + \cos\,(225\text{ deg.}) \right| = \left| 0.767 \sin\,(225\text{ deg.}) \right|$$

then

$$\left| K_2 - 0.707 \right| = \left| -0.542 \right|$$

and

$$K_2=0.165.$$

(c) If course C' were to be, say, 120 deg. from course B, the vertical radiation to be introduced must be of the same magnitude as in (b) but of opposite phase. Thus, θ_3 becomes 240 deg. and

$$\left| K_2+\cos 240\text{ deg.}\right| = \left| 0.767 \sin 240\text{ deg.}\right|$$
$$\left| K_2-0.5\right| = \left| -0.665\right|$$
$$K_2=-0.165.$$

(d) Suppose that an angle of 165 deg. between courses B and C' were desired. Course D is normally 180 deg. from B and is therefore the course to be used, with the proper amount of circular radiation introduced to obtain the desired angular shift. Courses D and A are symmetrically displaced about the 0-deg.–180-deg. axis and course A is 75 deg. from B. The angle of course D is therefore

$$\theta_4=300\text{ deg.}$$

and

$$\left| K_2+\cos 300\text{ deg.}\right| = \left| 0.767 \sin 300\text{ deg.}\right|$$
$$\left| K_2+0.50\right| = \left| -0.665\right|$$
$$K_2=0.165.$$

Note that for this value of K_2, course C is 90 deg. from B.

(e) Again, suppose that an angle of 195 deg. is desired between courses B and C'. Course D will again be used but circular radiation of negative sign is necessary.

$$\theta_4=315\text{ deg.}$$
$$\left| K_2+\cos 315\text{ deg.}\right| = \left| 0.767 \sin 315\text{ deg.}\right|$$
$$\left| K_2+0.707\right| = \left| -0.542\right|$$
$$K_2=-0.165.$$

Note that for this value of K_2, course C is 120 deg. from B.

V. METHOD D. EXTENSION OF METHOD B

It is frequently desirable to shift two of the four beacon courses from their 180-deg. relationship (viz., B and D) without disturbing the 180-deg. relationship between the other two beacon courses (A and C). This can be accomplished by the introduction of circular radiation *in equal amounts* of the carrier and sidebands transmitted by *both* amplifier branches in addition to their normal figure-of-eight radiation. Fig. 16 shows the beacon space pattern for a case of this type, and Fig. 17 shows the corresponding received polar pattern. The mathematical expressions corresponding to these patterns are given in (24) and (25), respectively.

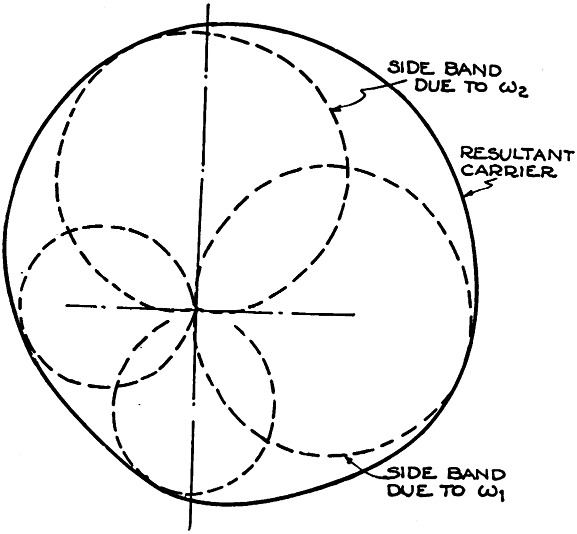

Fig. 16—Space pattern when circular radiation in equal amounts of the modulated waves transmitted by both amplifier branches is added to their normal figure-of-eight radiation.

$$e_p = K\left\{\begin{array}{l} E_0[\cos \omega t(K_1+\sin\theta)+\sin \omega t(K_2+\cos\theta)] \\ -\dfrac{E_1}{2}[\sin\,(\omega-\omega_1)t-\sin\,(\omega+\omega_1)t][K_1+\cos\theta] \\ +\dfrac{E_2}{2}[\cos\,(\omega-\omega_2)t-\cos\,(\omega+\omega_2)t][K_2+\cos\theta] \end{array}\right\} \qquad (24)$$

$$\text{where } K_1 = \frac{\text{amplitude of circular radiation due to } \omega_1}{\text{max. amplitude of figure-of-eight radiation due to } \omega_1}.$$

$$K_2 = \frac{\text{amplitude of circular radiation due to } \omega_2}{\text{max. amplitude of figure-of-eight radiation due to } \omega_2}$$

$$e_r = KK'E_0\left\{E_1 \sin \omega_1 t[K_1+\sin\theta]^2 + E_2 \sin \omega_2 t[K_2+\cos\theta]^2\right\}. \qquad (25)$$

A course will occur when (26) is satisfied.

$$|K_1+\sin\theta| = |K_2+\cos\theta|. \qquad (26)$$

In Figs. 16 and 17, $K_1 = K_2 = 0.2$

Applying method D, the minimum practicable angle that can be obtained between courses B and D is 145 deg. This angle can be made smaller, however, if the weakest course (C) is not to be used. For example, suppose that a radio range installed at Richmond, Va., is to serve the routes to Quantico, Va., Norfolk, Va., and Greensboro, N. C. The bearings of these cities from Richmond are, respectively, 1 deg., 126 deg., and 237 deg. These are plotted in full lines in Fig. 18, together with the proper beacon space pattern for serving the routes to these cities. The method of arriving at this pattern is of interest. Allowing an error of 0.5 deg., the courses to Norfolk, Va., and Greensboro, N. C., are equally displaced from the course to Quantico, Va. The bearings of two of the three courses may therefore be substituted successively in (26), and the values for K_1 and K_2 determined. Since the course directed on Quantico, which has a bearing of 1 deg., corresponds, obviously, to course A, Fig. 17, having a bearing of 45 deg., a correction factor of (+ 44 deg.) to the bearings of the two courses to be substituted in (26) must be applied. Thus

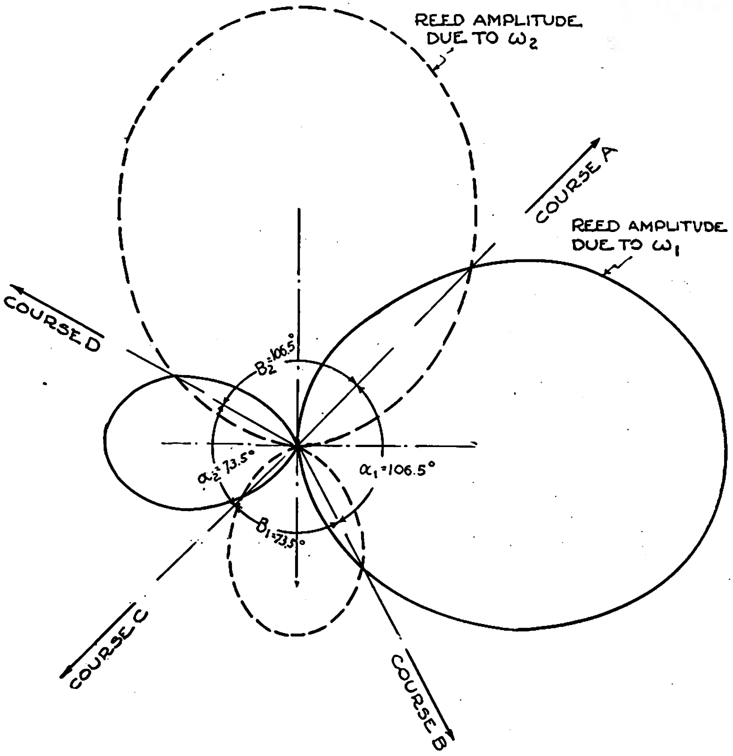
REED AMPLITUDE DUE TO ω_2
COURSE A
REED AMPLITUDE DUE TO ω_1
COURSE D
$\beta_2=106.5°$
$\alpha_2=73.5°$
$\alpha_1=106.5°$
$\beta_1=73.5°$
COURSE C
COURSE B

Fig. 17—Received polar pattern corresponding to Fig. 16.

$$|K_1+\sin(1\text{ deg.}+44\text{ deg.})| = |K_2+\cos(1\text{ deg.}+44\text{ deg.})|$$

$$|K_1+\sin(125.5\text{ deg.}+44\text{ deg.})| = |K_2+\cos(125.5\text{ deg.}+44\text{ deg.})|.$$

Solving these two equations simultaneously, we have

$$K_1=0.40$$

$$K_2=0.40.$$

A verification that these values are correct may be had by determining whether $|0.4+\sin(236.5\text{ deg.}+44\text{ deg.})| = |0.4+\cos(236.5\text{ deg.}+44\text{ deg.})|$. Solving $|0.582| = |-0.582|$, which is correct.

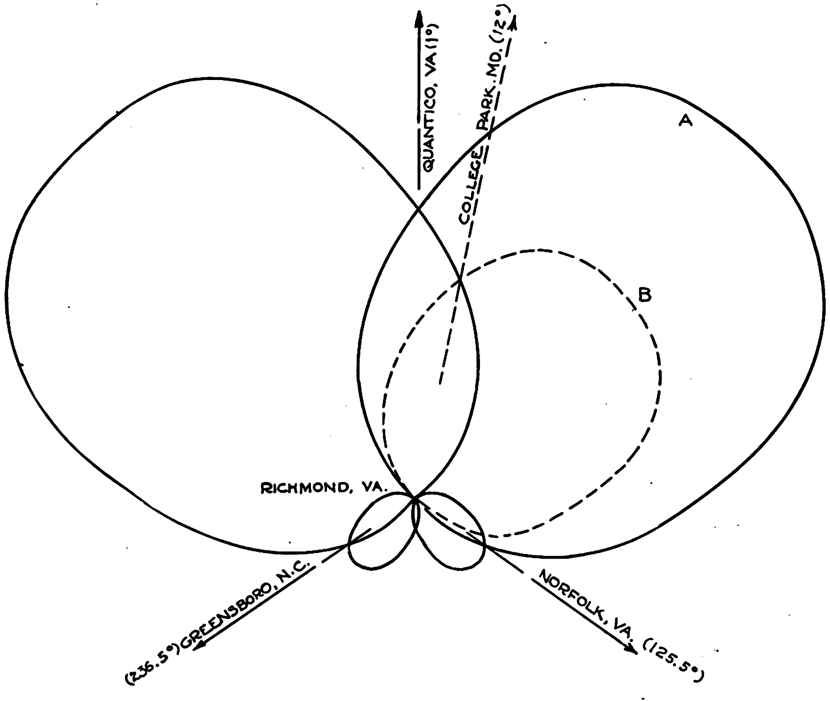

Fig. 18—Application of the method of course-shifting illustrated by Figs. 16 and 17 to actual airway routes.

Some difficulty arises in the application of this method due to the fact that no coupling should exist between amplifier branches of the beacon transmitter. Referring to Fig. 3, introducing circular radiation of the modulated waves transmitted by both amplifier branches of the system involves coupling the vertical antenna to suitable inductances inserted at A and at B. This at once introduces coupling between amplifiers and tends to destroy the beacon pattern. To overcome this difficulty the arrangement indicated in Fig. 19 must be employed. The inductances, S_1 and S_2, are inserted at A and at B as previously, but S_1 and S_2 are crossed at 90 deg. with each other. The secondary coils, R_1 and R_2, are mounted concentrically with S_1 and S_2 and are also at 90 deg. with each other. R_1 is connected in series with the tuned antenna circuit while R_2 is connected in a tuned circuit of constants identical with those of the antenna circuit. R_1 and R_2 may be made rotatable, making the coupling arrangement similar to the four-coil goniometer used for coupling the radio range transmitter to the loop antenna system. As was shown in the case of the four-coil goniometer, no coupling exists between amplifier branches, provided the loop antennas are in accurate tune.[4] The coupling between S_1 and S_2 by virtue of their mutual induction with R_1 is exactly neutralized by the coupling between S_1 and S_2 by way of R_2. This holds true for every angular position of the rotor system R_1 R_2.

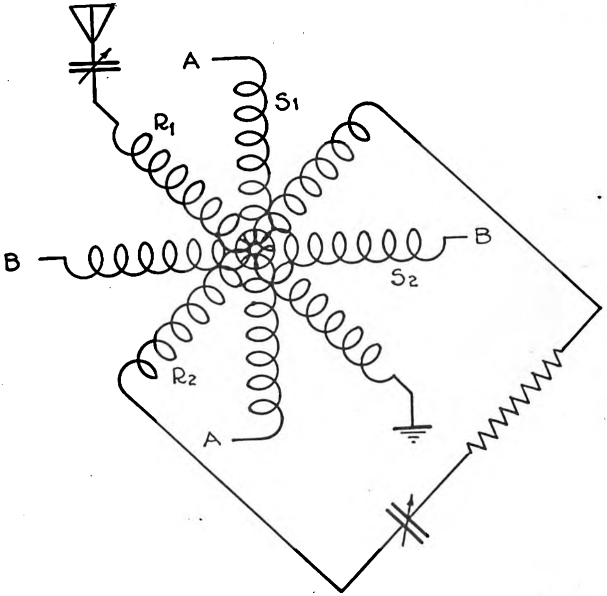

Fig. 19—Neutralizing arrangement to prevent intercoupling between the two amplifier branches even though these are coupled to the same vertical antenna.

The position of R_1 to obtain equal circular radiation of the modulated waves transmitted by the two amplifier branches is, of course, at 45 deg. witl S_1 and S_2. Changing the angular setting of R_1 results in a change in the relative amount of circular radiation for the two modulated waves. This provides an added feature of flexibility, since it makes it possible to shift the two sets of 180-deg. courses (*A*, *C* and *B*, *D*) from their 180-deg. relationship in varying amounts. This may best be illustrated by the results collected in Table II. The received polar diagrams shown in Figs. 13 and 17 correspond, respectively, to the 90-deg. and 45-deg. rotor settings in Table II.

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[4] J. H. Dellinger and H. Pratt, "Development of radio aids to air navigation," Proc. I. R. E., 16, 910–912; July, 1928.

TABLE II

Rotor Setting	K_2	K_1	Angle of Course *A*	Angle of Course *B*	Angle of Course *C*	Angle of Course *D*	Angle between Courses *D* and *A*	Angle between Courses *D* and *B*
0 deg.	0	0.283	33.5 deg.	146.5 deg.	236.5 deg.	303.5 deg.	203 deg.	157 deg.
14	0.068	0.274	36.6	149.0	233.4	301.0	196.8	152
18.5	0.090	0.268	37.5	149.8	232.5	300.2	195	150.4
26.5	0.126	0.253	40	150.5	230.0	299.5	190	149
45	0.200	0.200	45	151.4	225	298.6	180	147.2
63.5	0.253	0.126	50	150.5	220.0	299.5	170	149
71.5	0.268	0.090	52.5	149.8	217.5	300.2	165	150.4
76	0.274	0.068	53.4	149	216.6	301	163.2	152
90	0.283	0	56.5	146.5	213.5	303.5	157	157

VI. Method E. Combination of Methods D and A

A typical example will best explain the combination of methods D and A to give a flexible procedure for shifting the beacon courses so that they may coincide with the airways radiating from a given airport. Considering the Bureau's experimental station at College Park, Md., suppose that it is desired to direct courses on Hatboro, Pa., Norfolk, Va., Quantico, Va., and Bellefonte, Pa. The bearings of these cities from College Park are, respectively

Hatboro, Pa.	48.5 deg.
Norfolk, Va.	160.5 "
Quantico, Va.	219.5 "
Bellefonte, Pa.	343 "

These courses are plotted in Fig. 20, the angles between these courses being as indicated. The four courses can obviously be approximated by the method of amplitude reduction (method A) for shifting courses. Thus, draw line *AC* through the origin, equally displaced from the courses to Hatboro, Pa., and Quantico, Va. Also draw line *BD* through the origin, equally displaced from the courses to Norfolk, Va. and Bellefonte, Pa. The angle between these two lines is 62.2 deg. Placing $2 \tan^{-1} C_2 = 62.2$ deg. where C_2 is the reduction factor for the carrier and sideband in amplifier branch 2, we obtain

$$C_2 = 0.603.$$

It is now necessary to introduce sufficient circular radiation of both modulated waves to make courses *A*, *B*, *C*, and *D* coincide with the desired routes. The proper values for K_1 and K_2 may be determined by substituting the bearings of two of the desired courses in equation (27)

$$\left| K_1+\sin(\theta-12.9 \text{ deg.}) \right| = \left| K_2+0.603\cos(\theta-12.9 \text{ deg.}) \right| \quad (27)$$

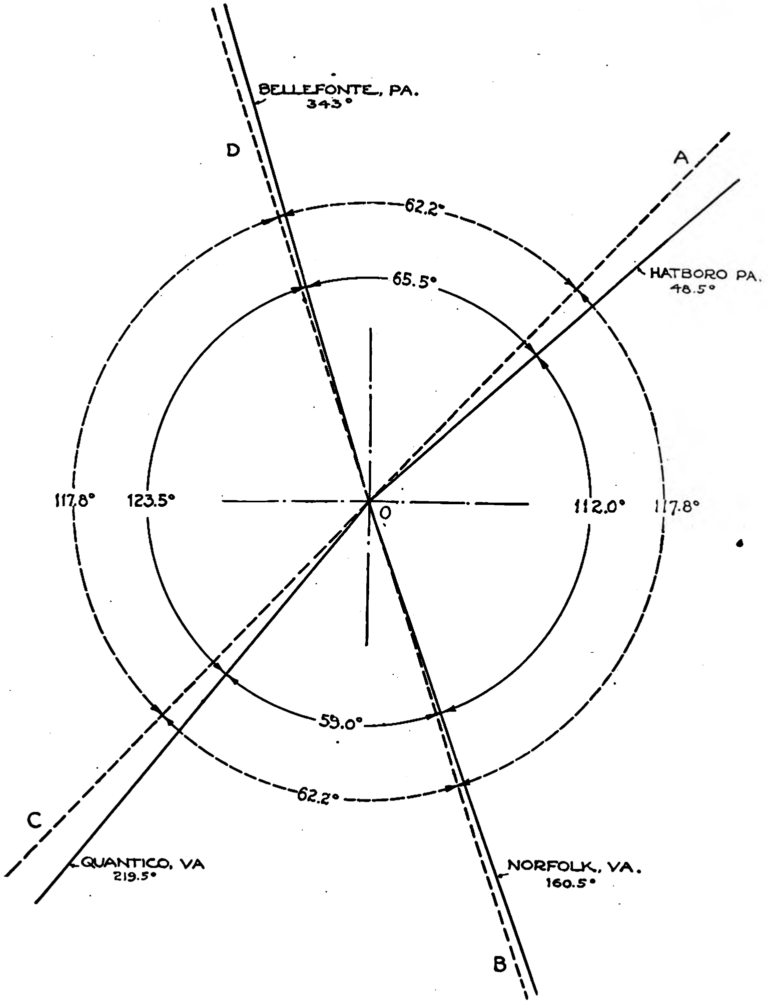
BELLEFONTE, PA.
343°
D
A
HATBORO PA.
48.5°
62.2°
65.5°
117.8°
123.5°
112.0°
117.8°
O
59.0°
62.2°
C
QUANTICO, VA
219.5°
NORFOLK, VA.
160.5°
B

Fig. 20—First approximation toward serving the four airway routes shown.

The correction angle 12.9 deg. is necessary to make courses *A* and *D* symmetrical about the 0-deg.–180-deg. axis (see Fig. 7). Substituting the bearings for the courses to Hatboro, Pa., and Norfolk, Va., in (27), we have

$$|K_1+\sin(48.5\text{ deg.}-12.9\text{ deg.})$$
$$=|K_2+0.603\cos(48.5\text{ deg.}-12.9\text{ deg.})|$$
$$|K_1+\sin(160.5\text{ deg.}-12.9\text{ deg.})|$$
$$=|K_2+0.603\cos(160.5\text{ deg.}-12.9\text{ deg.})|$$

Solving

$$K_1=-0.060$$
$$K_2=0.033.$$

The corresponding angular setting of R_1 (Fig. 19) is easily computed. The resultant space pattern is given in Fig. 21.

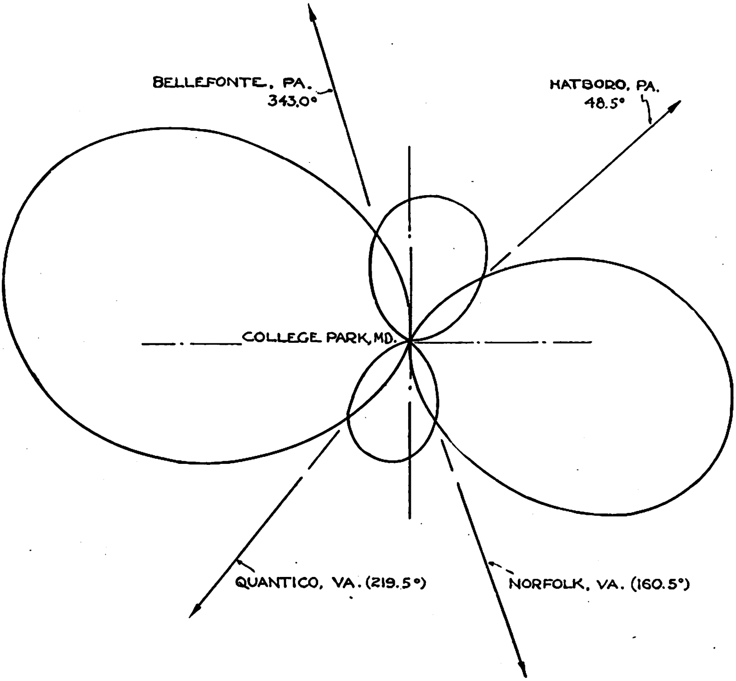

Fig. 21—Final polar pattern for serving the four routes shown simultaneously.

VII. Method F. Shifting the Courses by Changes at the Receiving Station

The methods for shifting the beacon courses as described above all require certain adjustments at the beacon transmitting station. Another possible method utilizes an adjustment of the receiving equipment aboard the airplane. This is accomplished by shunting a suitable resistance across the coil actuating one of the two reeds comprising the course indicator, thereby reducing the sensitivity of that reed. The course, as determined by equality of reed deflections, is therefore shifted from the true equisignal zone of the beacon in the direction of the shunted reed. The same effect is thus accomplished as by shifting the courses at the beacon.

This method may be used alone, that is, with the beacon normally adjusted for four courses at 90 deg. to each other, or it may be used in conjunction with any of the methods outlined above. In the former case, a course shift of 15 deg. on either side of the four-beacon courses can be obtained. In the latter case, the degree of course shift possible (at the same time fulfilling the criteria set up under method A), depends upon the polar pattern employed.

An example of the latter case may be seen by reference to Fig. 18. A course directed on College Park, Md., in addition to the three courses already indicated, is desired. The pilot may obtain this course in preference to the one directed on Quantico, Va., by shunting the proper reed by a suitable resistance, thereby obtaining the dotted line pattern *B* on that reed in place of the full line pattern *A* obtained with the reed unshunted.

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RADIO IN AERONAUTICS—ITS TECHNICAL STATUS AND THE ORGANIZATION FOR ITS APPLICATION IN GERMANY*

By

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Summary—*The paper gives a review of the present status of radio in aeronautics in Germany. Not only does it describe the apparatus used at the present time in German aircraft communications, but it also gives results of measurements on the fundamentals of radio in aeronautics, which have been made by the Deutschen Versuchsanstalt für Luftfahrt E. V. Further, it indicates the trend that German aircraft radio will probably follow in the next few years.*

It takes up the subjects of long-wave sets, sets for short- and ultra-short waves, navigation apparatus for aircraft using radio waves, and airship radio sets. Finally, the German organization for the application of radio in aeronautics is described.

I. Introduction

PEOPLE who are not in contact with radio in aeronautics are frequently of the opinion that it is sufficient to build given transmitting and receiving apparatus for radio aircraft communication to meet its special problems. The use of radio apparatus in aircraft imposes on its design the requirements of minimum weight and bulk. In order to meet this condition, the construction of such apparatus is made difficult and expensive, because the utilization of space must be carried to such a point that installation frequently becomes exceedingly difficult. In addition to these requirements of weight and bulk that have been placed on aircraft radio, there is an additional requirement that the apparatus must be so designed and built that, considered purely mechanically, it will withstand vibrations during flight. This includes mounting the tubes on springs, special construction of the variable condensers in order to prevent displacement by vibration, and many other things. But it is even more important that the apparatus be electrically adapted to its special problems as far as possible. This is true of the transmitting as well as receiving apparatus.

Because of the extremely great restriction in weight, the most suitable selection of the aircraft radio wave, with due consideration of the problem of propagation, is much more important than in immobile stations. The determining factor for propagation is the so-called attenuation constant, the knowledge of which is necessary over land and sea. Over the sea, the propagation process with long waves has been sufficiently covered by the researches of Austin and many other investigators.[1] The work[2] done by Deutschen Versuchsanstalt für Luftfahrt E. V. is of value for the attenuation constants over land in relation to the wavelength, and the data can be used in planning aeronautical radio connections.

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In the short-wave region, propagation cannot be expressed by formulas. Here the knowledge of the so-called "dead zones" or, to use a better term, "fading zones," is particularly important.

In connection with the transmitter, the current supply brings up special problems. In aircraft radio stations, the energy is frequently taken from a propeller, so that the question of constant voltage in spite of variable speed of the aircraft in relation to the surrounding air is not a simple technical problem.

Also, the question whether pure undamped or audio-modulated undamped waves are the most suitable type for airplane transmission is also important. Here this question must be considered separately for transmitting and receiving. In telephony, for example, there are special difficulties in transmitting from airplanes because the microphone is affected not only by the acoustic speech waves, but also by all general noises.

With short waves other questions come to the front in connection with transmission, such as the suitability of quartz control.

A special question is raised by the shape of antenna suitable for aircraft. First, the antenna must be so shaped that it will cause, aerodynamically, only a slight reduction in the flying speed. Then, in special cases, such as low flying, it should interfere with the flight as little as possible. Finally, the exact knowledge of the radiation resistance and of the effective height of the airplane antenna is more important than in land stations, as these two magnitudes determine the range, and the necessary ranges can be obtained with low powers only with the correct shape of antenna. In aviation, no shape of antenna should be used if we do not know its aerodynamic action, and especially its radiation efficiency and its effective height. This requirement had not been met up to a short time ago. In addition to the requirements placed on the electrical properties of the antenna, it must also be ascertained that the shape of the antenna does not endanger the aircraft. This might be the case if a fixed antenna breaks as a result of overstressing, and parts of it get into the power plant and control surfaces.

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[1] Sacklowski, "The propagation of electromagnetic waves," Berlin, *Weidmannsche Buchhandlung*, 1928.

[2] F. Eisner, H. Fassbender, and G. Kurlbaum, "Investigations on the attenuation of electromagnetic waves and the ranges of radio stations in the wave region from 300 to 2000 m," (to appear shortly).

There are also a number of fundamental matters on the receiving side, in which aircraft radio apparatus differs from that in immobile radio stations. For example it must first be decided whether telephone or telegraph signals are better suited for reception in an airplane. This is tested by a type of articulation test[3] similar to that generally employed in cable work at the present time. The data applying in stationary radio communication concerning the relation between the wireless telegraph and telephone ranges do not hold at all in aeronautic radio; in this case the ratio favors telegraphy. There are other general

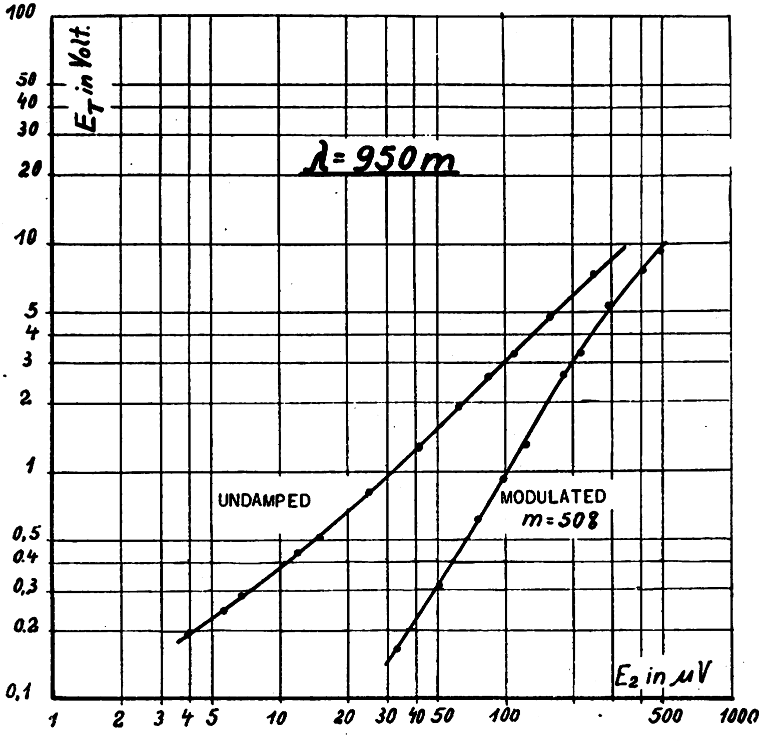

Fig. 1—Output voltage E_T as a function of input voltage E_2.

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[3] F. Eisner, "Use of articulation measurements in radiotelegraphy (from experiments made with G. Kurlbaum). *Zeitschrift für technische Physik* (to appear shortly).

questions on reception in connection with this, especially the receiving field intensity that is necessary in a given airplane, as compared with the field intensity necessary for equivalent reception on the ground.

The list of questions in which aircraft radio differs essentially from that of ordinary radio engineering could be increased still further, but the points that have been mentioned above are sufficient to show that the old view is incorrect; the problems of aircraft radio are not merely special problems of design.

II. LONG-WAVE APPARATUS

The apparatus most widely used in German aircraft at the present time is the so-called 70-watt set of Telefunken, type 262 F. The method of rating the power of a radio station is not absolutely fixed, and it is not sufficient to characterize the set according to one of the many possible definitions of power; the product of transmitting current and effective height of aerial is important for operation. But this data depends on the aerial used, and which the manufacturing firm cannot always control. Aside from this, a set is best characterized by the power that can be delivered between the terminals: antenna —ground. According to the measurements of the Deutschen Versuchsanstalt für Luftfahrt E. V.[4] we get for the Telefunken set:

TABLE I

POWER DATA FOR TELEFUNKEN TYPE 262F AIRCRAFT TRANSMITTER (70 WATT).

Wavelength m	Frequency kc	N_a watt	N_{ak} watt	N_{ant} watt	$J_{ant}.h_{eff}$
450	667	193	64	28	19.5
650	462	193	73	24	18.8
950	316	193	81	23	19.8
1350	222	193	90	14	16.3

N_a power taken at the anode
N_{ak} total power in the antenna circuit including the tuning apparatus
N_{ant} power in the antenna wire
h_{eff} effective height of the usual trailing antenna in Germany

The wavelength range of the Telefunken set is from 300 to 1350 m. It contains a separately excited tube transmitter. The master oscillator has a 3-point regenerative connection in which the wavelength is changed by adjusting a variable condenser and variometer, both of which are on the same shaft. The grid circuit of the power tube is inductively coupled to the master oscillator. In the plate circuit of the power tube are the coupling coils for the antenna, likewise coupled inductively. The couplings are variable in steps together with the tuning of the antenna circuit. A variometer permits fine tuning of the antenna circuit. Power tubes and control tubes are those of the RS31 and RS231 type, respectively, with a filament potential of 14 volts, a filament current of 5 amperes, and a plate potential of 1500 volts. Although the control tubes could be considerably smaller, two tubes of the same

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[4] F. Eisner, H. Fassbender, and G. Kurlbaum, "Power and radiation measurements on airplane and ground stations," 99th Report of the Deutschen Versuchsanstalt für Luftfahrt E. V., Berlin-Adlershof, *Jahrbuch der drahtlosen Telegrafie und Telefonie,* 31, No. 4/5, 1928.

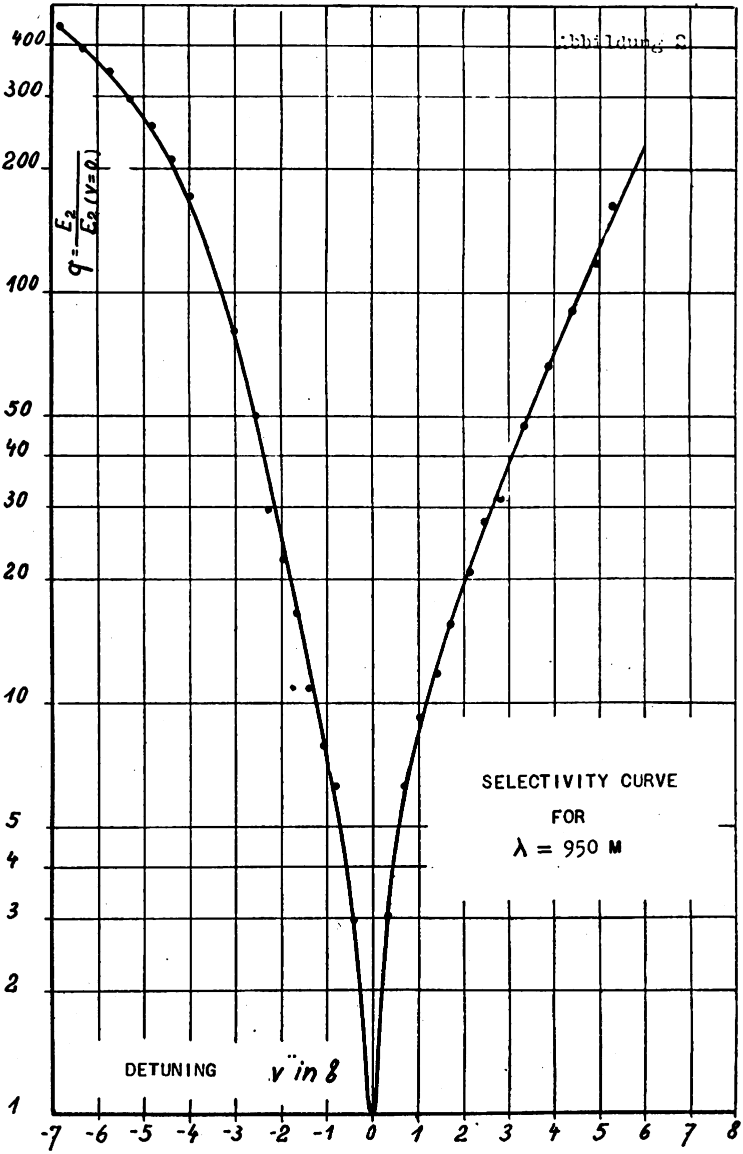

Fig. 2

size have been chosen in order to use the same voltage for both tubes and to simplify the storage of spare parts. The transmitter permits the use of pure undamped waves and is also arranged for audio-modulated waves and for telephony. Modulation is carried out by varying the grid current. The modulation tube is heated by an a-c voltage of 80 cycles, obtained from the generator through slip rings.

The transmitter is enclosed in a mahogany box, and the adjusting knobs can be protected by a cover. All cables are plugged into the lower wall of the box. The front wall carries all the tuning knobs and the knobs for changing to "off, transmitting, receiving," and also the switches for the different types of transmitting. Recently the transmitter has been provided with a listening-in device which consists of a small audio-frequency tube transmitter whose plate potential is supplied by the high-frequency field in the vicinity of the transmitter and the output circuit of which is in series with the output circuit of the receiver. The maximum antenna current with pure undamped transmission is 2.3 amperes.

The receiver is placed on the transmitter, and both have the same base area. All control knobs and cable leads are on the front wall. The Telefunken set has a 3-tube secondary receiver with single-stage high-frequency amplification, one detector tube and one tube for low-frequency amplification. At the present time tubes of the RE1 type are used, which are double-grid tungsten-filament tubes with a filament potential of 4 volts and a plate potential of 50 volts. The filaments of the tubes are connected in series and are heated direct from the low-voltage d-c winding of the generator through a common resistance. The plate voltage is taken from the high-voltage winding for the transmitter by a potentiometer (voltage divider). The wave range of the receiver is also from 300 to 1350 m. An aperiodic antenna is used for receiving. The coupling is adjustable in three stages. The two tuned circuits are operated by a common drive. Adjustable regeneration is provided. The telephone receiver is sewn into a cap.

The sensitivity of this receiver with a 950-m wave can be seen from Fig. 1, which shows the output voltage E_T in relation to the input voltage E_2 for pure undamped reception and for audio-modulated undamped reception, according to measurements of the Deutschen Versuchsanstalt für Luftfahrt E. V. Fig. 2 shows a selectivity curve for the same receiver with the same wavelength, in which for a constant output voltage depending on the percentage detuning V, a value q has been plotted, which gives the ratio of the input voltage required for constant output to that corresponding to resonance.

Transmitter and receiver are supplied with energy from a generator driven by a regulator-type wind-propeller. The speed is 4800 r.p.m.

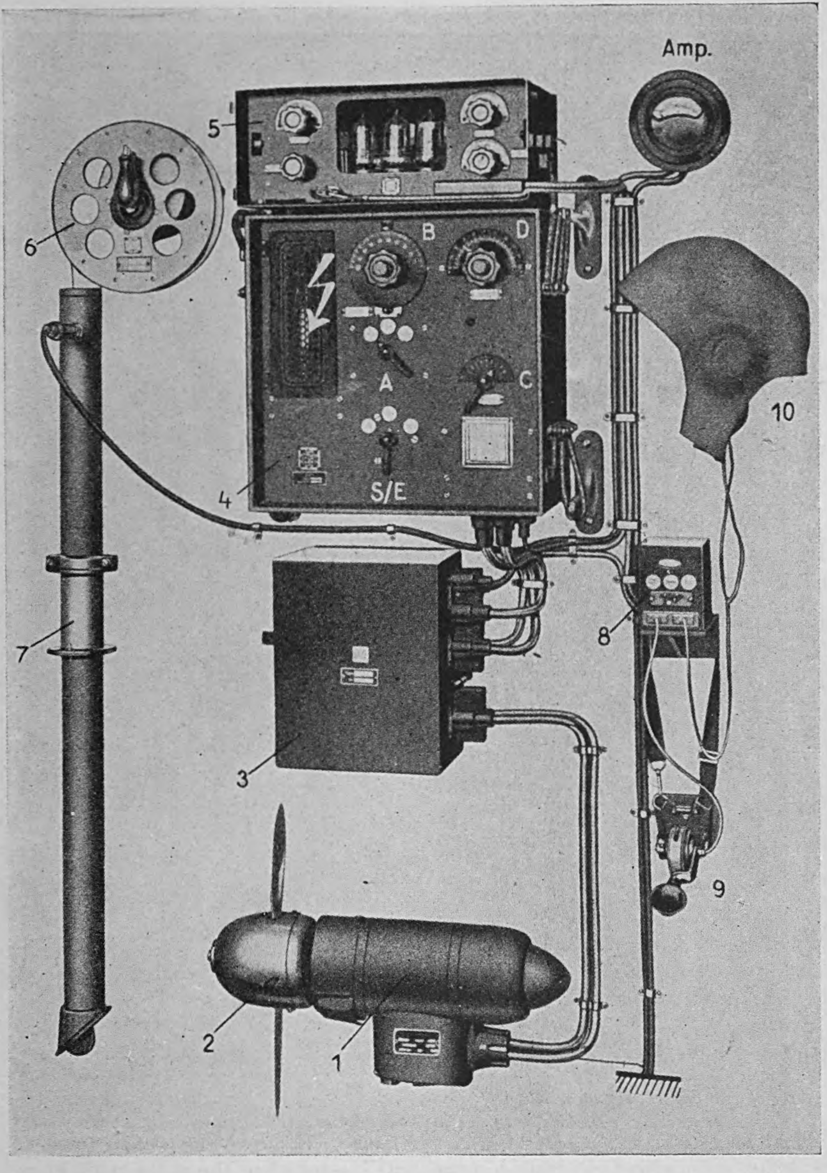
Amp.
5
B
D
6
A
C
10
4
S/E
8
7
3
9
2
1

Fig. 3—Separate parts of Telefunken transmitter 262F.

The generator delivers 14 volts direct current for heating the transmitter and receiver tubes, 2000 volts direct current for the plate potential, and alternating current at about 9 volts at 80 cycles, for heating the modulation tubes. According to measurements of the

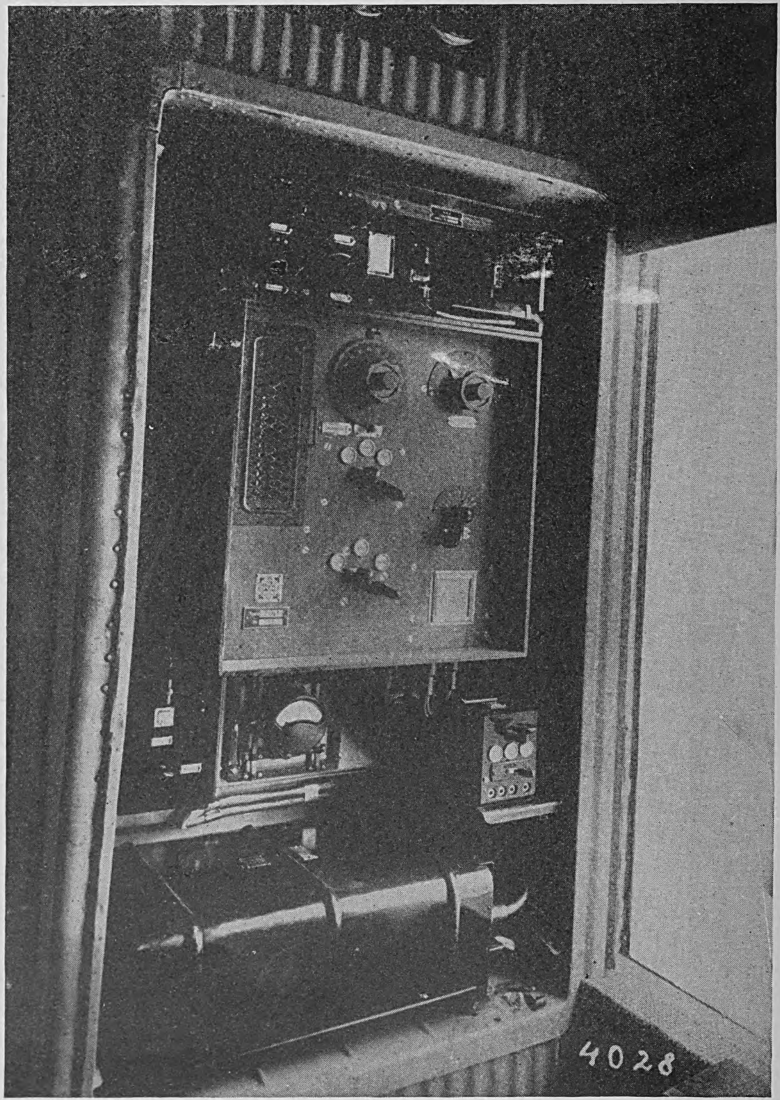

Fig. 4—Telefunken station, type 262F, installed in the radio cabin of a Junkers G31.

Deutschen Versuchsanstalt für Luftfahrt E. V.[5] the propeller absorbs 1100 watts of power from the wind. The latest type of generator has a waterproof casing. In order to reduce the commutator noise, the high-tension winding is connected in parallel to a large condenser, and chokes are placed in the line for the filament current of the receiver. The chokes are all placed in one sheet metal case, which also contains the resistances for balancing the transmitter and receiver filament current and the potentiometer.

The antenna is a bronze braid 70 m long, which is weighted at the end by a "fish" weighing 0.5 kg; it is put out through an antenna well, generally through the bottom of the airplane. The exact determination of the effective heights of airplane antenna offers considerable difficulties. The Deutschen Versuchsanstalt für Luftfahrt E. V. has found 8.5 m for the trailing antenna 70 m long.[4] The antenna well is a "pertinax" (bakelized paper) tube about two meters long which has recently been given a stream-line profile. The aerial is wound on a reel and runs out automatically when operation starts. A centrifugal brake controls the speed. Turning the handle on the reel operates a stopping device.

The key is separated from the apparatus, and controls the grid current. The Telefunken set weighs about 49.0 kg (100 lbs.). Its certain range with telegraphy using pure undamped signals and with normal atmospheric conditions is 500 km (310 miles) for transmitting from airplane and receiving from a 1-kw airport transmitter. Fig. 3 shows the individual parts, and Fig. 4 shows a set installed in the radio cabin of a G31 Junkers plane.

The C. Lorenz A. G., type Serf 01 V 28, airplane radio set is externally similar to the Telefunken set. It also consists of transmitter and receiver, but has in addition a so-called panel, on which are placed all switches and fuses. It differs from the Telefunken type in many interesting ways. The weight of the set is about 48.2 kg and its rated power is 100 to 120 watts. Measurements by the Deutschen Versuchsanstalt für Luftfahrt E. V. on this set gave the figures in the following table.

TABLE II

POWER DATA FOR LORENZ, TYPE SERF 01V28, 100-WATT AIRPLANE TRANSMITTER.

Wavelength m	Frequency kc	N_a watt	N_{ak} watt	N_{ant} watt	$J_{\text{ant}} \cdot h_{\text{eff}}$
650	462	640	198	71	32.3
950	316	476	196	52	28.9
1350	222	420	132	29	23.7

The letters have the same meaning as in Table I.

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[5] W. Brintzinger, "Driving electric generators by wind," 101st Report of the Deutschen Versuchsanstalt für Luftfahrt E. V., Berlin-Adlershof, *Luftfahrtforschung*, 1, No. 4, 1928.

The Lorenz transmitter is a self-excited intermediate circuit transmitter. Its wavelength range is from 550 to 1350 m. The wavelength adjustment is done in steps by plugging-in. Fine adjustment is furnished by a variometer and in addition, the antenna coupling, antenna tuning and also grid and plate coupling of the transmitter tube can be changed in steps. A variometer is provided for fine tuning of the antenna. The filament current of the transmitter tube is measured by an ammeter and can be adjusted by a sliding resistance. All cable connections are at the lower wall of the box. The maximum antenna current is about 3.3 amperes and is measured by a hot-wire ammeter connected in the ground wire. The transmitter is adjusted for pure undamped and audio-modulated undamped transmission as well as for telephony. The modulation is magnetic.

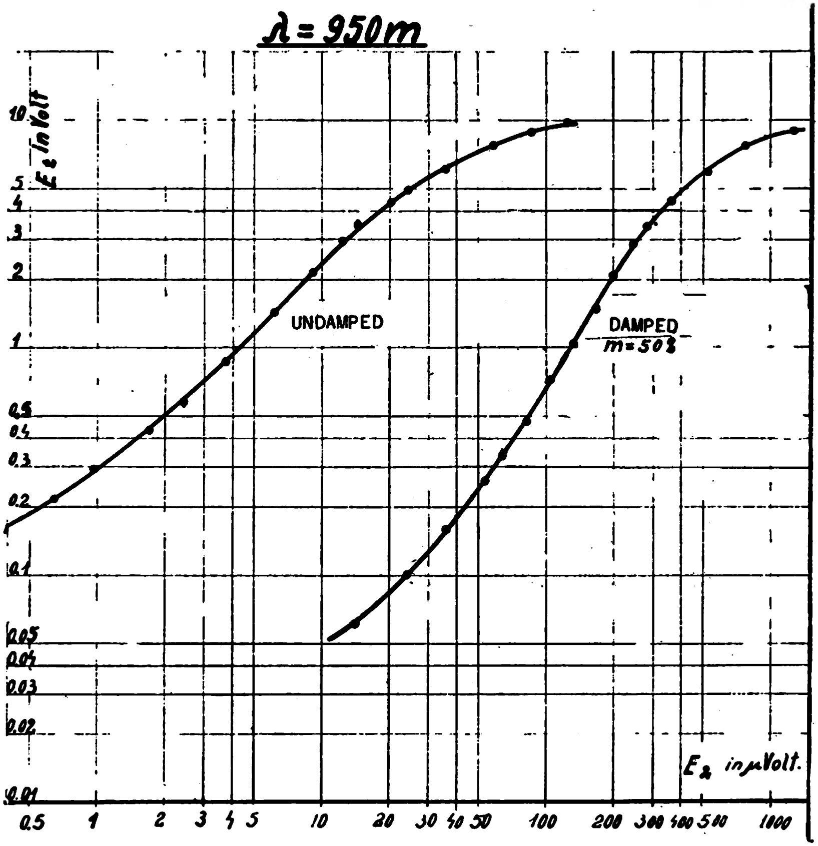

Fig. 5—Output voltage E_T as a function of input voltage E_2.

The receiver has four tubes of the RE11 type, which are employed as high-frequency amplifier, detector and 2-stage audio-frequency amplifiers. In this receiver the antenna is tuned, and for this purpose the tuning device of the transmitter is used. If transmission and reception are on different waves, it seems inconvenient to change the antenna tuning for each shift, but the receiver does not have a sufficient range with an untuned antenna.

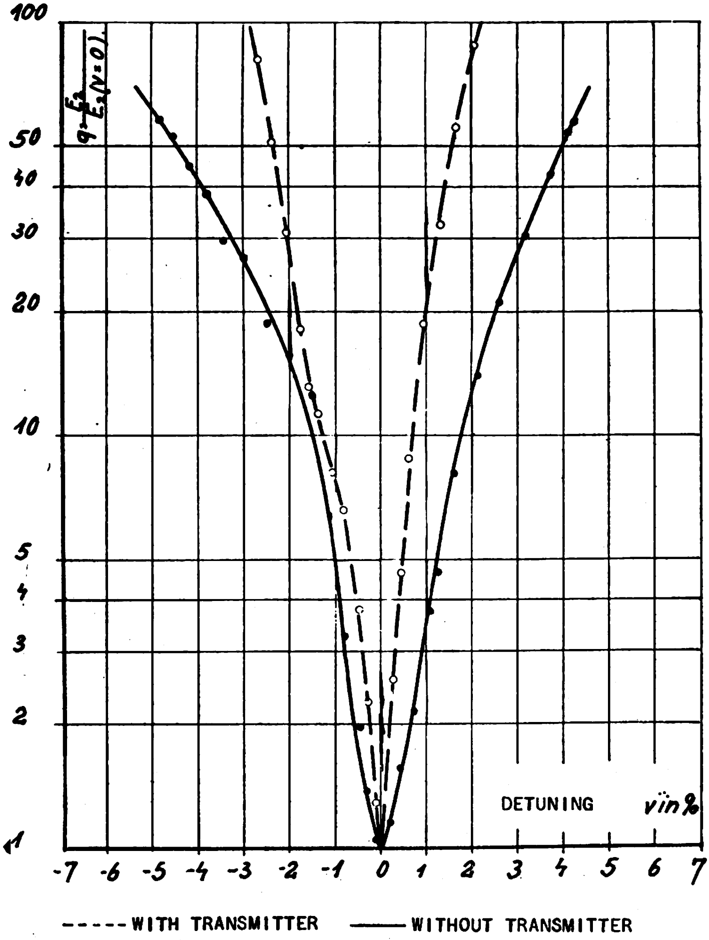

Fig. 6—Selectivity curve for $\lambda = 950$ m.

In the receiver there are two tuned circuits which are adjusted by turning two knobs. Fine tuning is provided for the second circuit. The regenerative coupling is variable. The receiver in the airplane can receive from the 1-kw airport transmitter at a distance of about

600 km (375 miles). The sensitivity of this receiver on the 950-m wave can be seen from Fig. 5, which shows the output voltage E_T as dependent upon the input voltage E_2 for pure undamped and for audio-modulated undamped reception, according to measurements of the Deutschen Versuchsanstalt für Luftfahrt E. V. The selectivity curve ($\lambda = 950$ m) for the Lorenz receiver is shown in Fig. 6.

The receiver is heated by a 24-v direct-current winding on the generator, and the plate voltage is taken from a special 220-v direct-current winding on the generator. The receiver remains connected during transmission, so that the transmitted signals can be heard in the telephone.

In the Lorenz set the receiver is attached to the transmitter. All cables are fastened to the underside by means of plugs. The current for this set is supplied by a generator driven by a regulator propeller.

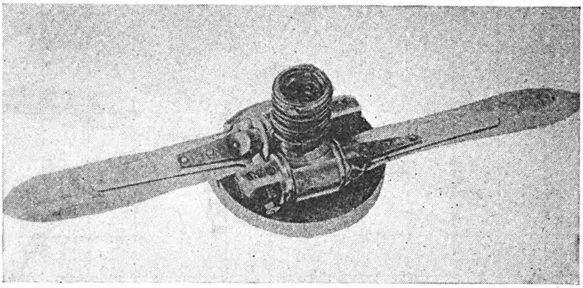

Fig. 7—Telefunken adjustable propeller type ED 2574.

Fig. 7 shows the propeller for this set made by Seppeler-Telefunken. Each of the two blades can be turned about its longitudinal axis. Regulation takes place as follows: the centrifugal force acts on weights which are placed eccentric to this longitudinal axis, so that with increased speed the blades are given a smaller angle of attack. The propeller shown in the figure supplies the necessary counterforce by cable tension. The generator supplies 165 w at 24 v for heating the transmitter and receiver tubes, 370 w at 2000 v for the plate circuit of the transmitter, and 15 w at 220 v for the plate circuit of the receiver. The generator also supplies 5 va at 1000 cycles for modulation in audio-modulated transmission. It should be mentioned in connection with the generator that it is provided with a cowling which gives the least possible air resistance. In addition, the shaft is hollow for better cooling, so that the inside of the machine, which heats up more, is cooled most.

A condenser is connected to the plate-voltage winding of the transmitter to remove harmonics, and filament and plate currents for the receiver are sent through a series of chokes for the same reason.

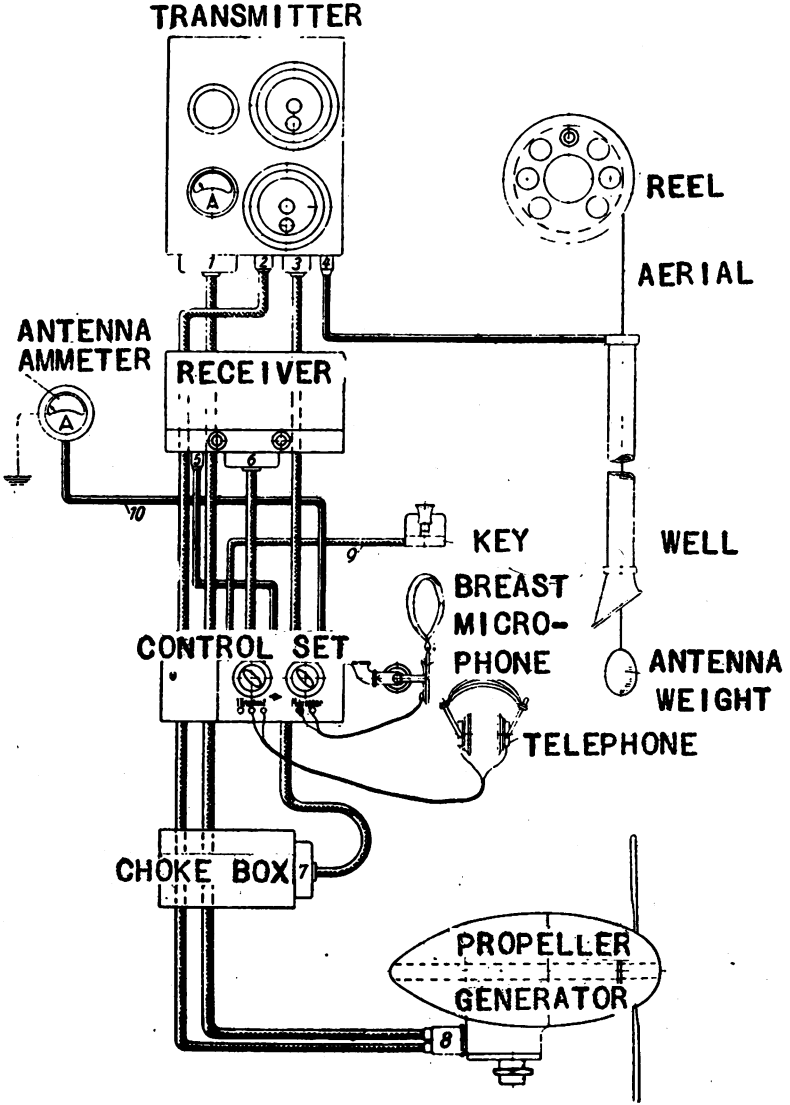
TRANSMITTER
1
2
3
4
REEL
AERIAL
ANTENNA AMMETER
A
RECEIVER
5
6
10
KEY
9
WELL
BREAST MICRO-PHONE
CONTROL SET
ANTENNA WEIGHT
TELEPHONE
CHOKE BOX
7
PROPELLER GENERATOR
8

Fig. 8—Lorenz set, type SERF 01V28. Separate parts.

The Lorenz set also uses as antenna a bronze braid 70 m long, which is wound on a hand-operated reel equipped with a brake. The reel has a stop which is activated by turning the handle. Here also, the well is a

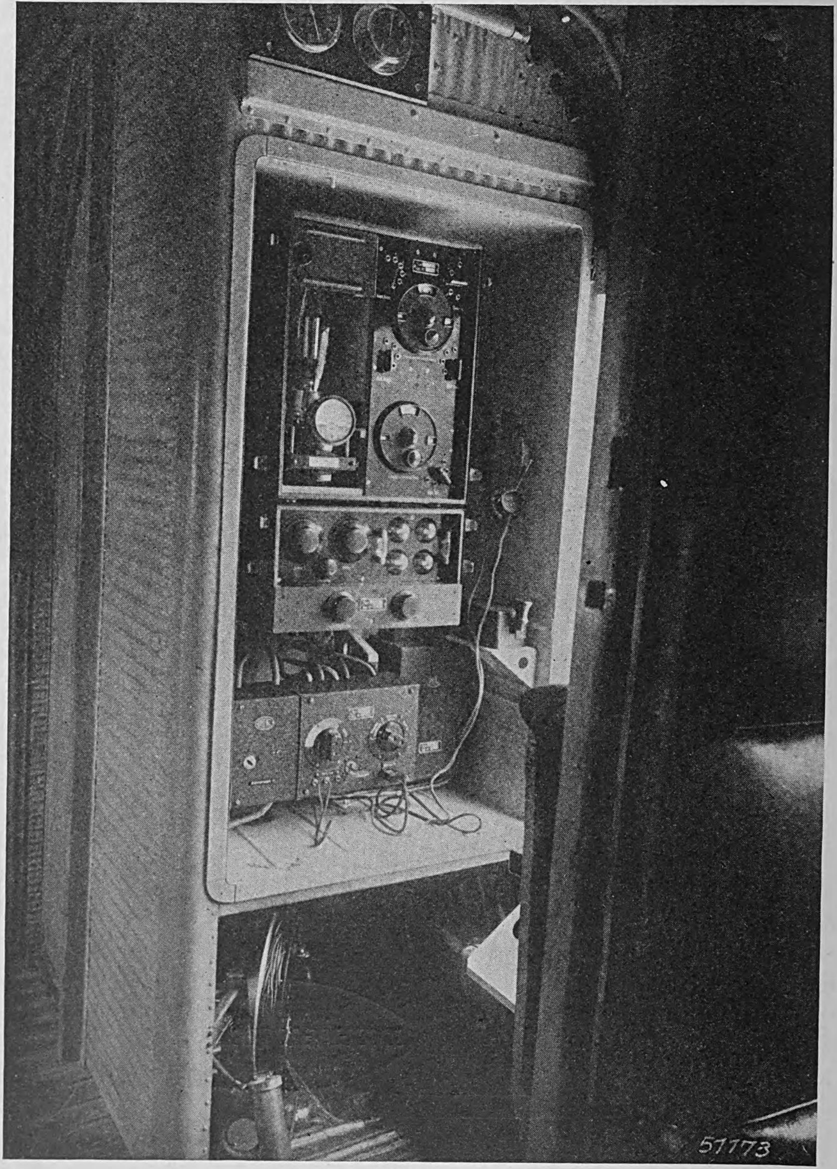

Fig. 9—Lorenz set, type SERF 01V28, installed in the radio cabin of a Junkers G31.

pertinax tube about 2 m long, which has such a wide internal diameter that, if an antenna weight is lost, the metal tube can be removed during flight, in order to replace the weight by another.

The Lorenz set and a 1-kw airport set can maintain reliable communication with pure undamped operation and normal atmospheric conditions, for 600 km (375 miles).

Fig. 8 shows an assembly of the parts of this set, and Fig. 9 shows an installation in a G31 Junkers plane belonging to Lufthansa.

To summarize, we can say, in regard to the two airplane sets in common use at the present time in German aircraft radio, that they fully meet the requirements placed on the apparatus by this special type of operation. The sets possibly give too much, for in communication within Germany, distances of more than 500 to 600 km (315–375 miles) are never required. Use is rarely made of the possibility of audio-modulated transmission. At most, it will be used for communication in the vicinity of the coast, with ship and coast stations which as yet have no heterodyne receivers. Telephony is not used in Germany because thorough investigations have shown that, especially for reception in airplanes, this will require such great field intensities that the range will be restricted to very short distances. The Deutschen Versuchsanstalt für Luftfahrt E. V. took up experimentally the question of the extent to which telephone reception is poorer in airplanes

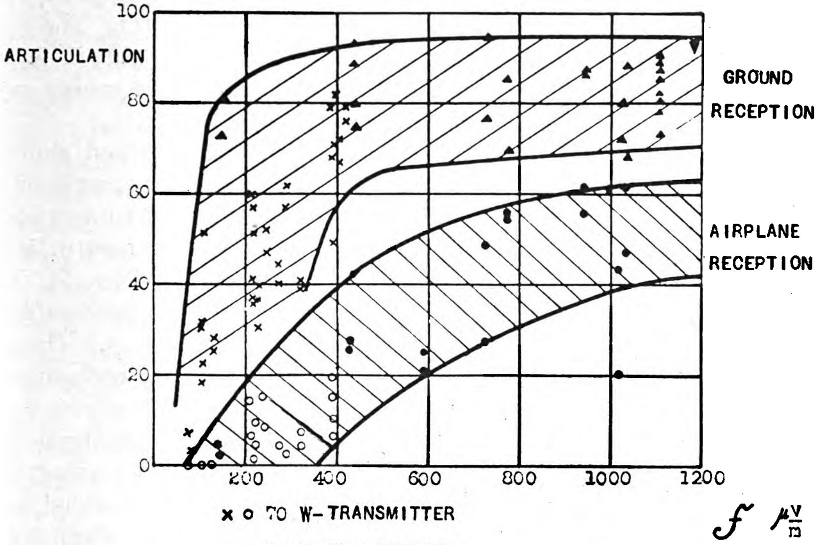

Fig. 10—Articulation in relation to receiving field intensity. Receiver (1).

than on the ground, using the same type of receiver and the same reception field intensity.[3] Fig. 10 shows for aircraft and ground reception the percentage articulation as a function of the necessary field intensity. The disadvantages for the development of telephone operation also consist in the fact that more errors can occur in transmission than with telegraphy. The disadvantage of having a skilled operator on board for telegraph operation is more tolerable.

Work is now in process to simplify the two sets that have been described and which have proved to be good. It is intended to change the present form of the sets in such a way that the arrangements for audio-modulated and speech-modulated transmission can be omitted. This gives simplified construction and less weight, and in addition requires fewer repairs and a smaller supply of spare parts. The receiver is to be simplified by using low-drain (sparröhren) tubes instead of tungsten filament tubes. This will greatly reduce the energy required by the set so that its generator can be made lighter.

In addition to the long-wave sets that have been described, short-wave sets have been developed, which are being tested as sample sets by Lufthansa at the present time. These were made by Telefunken and Lorenz after the technical and scientific bases for their construction had been given by Deutschen Versuchsanstalt für Luftfahrt E. V. These sets are so small that they can be placed in the pilot-cabin. The weight of the set, ready to operate, is about 20 kg (40 lbs.). Operation is so simple that it can be done by the second pilot or radio operator. The sets are adjusted for definite wavelengths and provision is made for tuning the antenna to these wavelengths. Fluctuations of the antenna capacity have just as little effect on the wavelengths as in the large sets. Only pure undamped telegraphy is provided for the transmitter. The receiver has a wavelength range from about 850 to 950 m and only tuning and regenerative-coupling adjustments. The set is supplied by a 12-v, 18 ampere-hour battery. The antenna is a trailing wire 70 m long. The transmitting antenna current is 1 ampere. The reliable range is 200 to 250 km (125–160 miles) over land.

The small sets are intended to fill an important gap in aircraft radio engineering. They make reliable aircraft radio communication with the numerous airport radio stations in Germany possible at any point. They make it possible for the medium and small airplanes for which they are designed, to orient themselves above the storms, and to get bearings. Even after forced landings, communication can be carried on from the airplane, if the aerial is removed and fastened temporarily in some manner.

The ground stations necessary in German aircraft radio service generally consist of a transmitting and a receiving set, but there are also airports that have only radio receiving sets, and some landing fields are connected with the other airports only by cable. The airport transmitters are most generally tube transmitters with a power of 0.3 to 1.5 kw in the antenna circuit. Only the Lindenberg transmitting station, which takes a part of the radio traffic of Berlin when necessary, has a high-frequency machine-transmitter supplied by the Lorenz firm, with a power of 5 kw; another Lorenz machine-transmitter is also installed in the transmitting station of the Munich airport. The wavelength range of the machine transmitter is from 800 to 1900 m.

The tube transmitters built by the Lorenz firm are separately excited and have intermediate circuits. The current supply in most cases is from 500-cycle machines whose current is rectified by thermionic valves. Some transmitters are also equipped with direct-current high-voltage machines. The wavelength range of these transmitters is from 500 to 2400 m.

The transmitters made by Telefunken are self-excited and have an intermediate circuit. Here also the high potential is produced by rectified alternating current of 500 cycles. Direct-current machines are used for cathode heating.

Almost all airport transmitters are at some distance from the flying field, generally 1 to 2 km, so that the aerial towers do not interfere with flying, and also in order to avoid interference in reception by its own transmitting. They are remote-controlled from the receiving room. The aerial towers are self-supporting iron masts. A few transmitting stations, such as Berlin, in which ground could not be obtained in the vicinity of the flying field, have been built on airport territory and have self-supporting wooden masts. The antennas are generally of a 3 or 4 wire, *T* or *L*, type. Grounding is done by radiating buried copper wires. The transmitters are in brick houses which, in addition to the operating and engine rooms, also contain a dwelling for the chief of the transmitting station or for the radio operator. Fig. 11 shows a view of the Nurnberg transmitting station, which can be regarded as characteristic of the airport transmitting stations.

The receiving sets belonging to the airport radio stations are always in the immediate vicinity of the airport. In general, they are placed in the administration building of the company operating the airport and near the weather bureau office.

A receiving station generally has three receivers. Most of the receivers are the Telefunken type E 266, consisting of detector and two audio-frequency amplifier stages which, in part, operate with separate heterodyne oscillators. Modern receivers such as DEBEG type E364S, or the Lorenz 6-tube neutrodyne receiver, type ERNC II

627, are being introduced more and more. As the international weather service uses very long waves (up to 12,000 m) each station is provided with a receiver whose wavelength range can be increased to the longest waves by changing coils. The E 266 is very suitable for this.

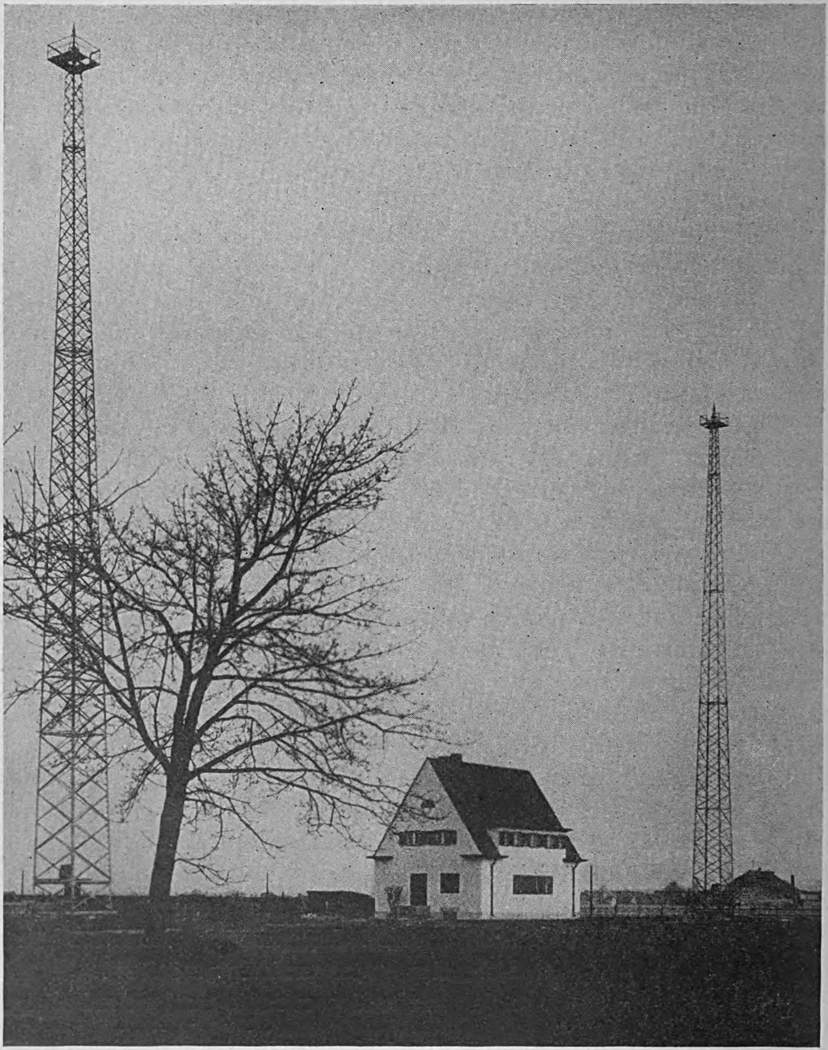

Fig. 11—Nurnberg airport transmitting station.

The receiving aerials are ordinary low steel pipe masts and the grounds are buried conductors connected to the water pipe.

It is noteworthy that, as a result of uninterrupted operation, the wear of the plate batteries was so great that all receiving stations had to be equipped with net plates (Netzanoden). Storage batteries are used for heating.

III. SHORT-WAVE APPARATUS

The greatest value of aircraft communication lies in bridging great distances. The range that can be attained with long-wave sets does not meet the justified demands when constant communication with the home station is to be maintained on long flights. Here, distances of several thousand km come into consideration.

The trailing antenna (70 m long) used in the present sets hinders operation of the airplane, particularly in poor weather, so that for a long time there has been a demand for fixed antennas. But, as the distances available on the body of the airplane are very small as compared with the wavelengths customary in aircraft communication at the present time, fixed antennas have a decidedly low efficiency in the present sets, so that they have only very short ranges.

Further, the problem of so building a set that it can be used during flight as well as after landing on the ground, or after alighting on the sea, is very difficult with long-wave sets. But this problem is one of considerable importance, as wireless communication is particularly essential after a forced landing. Even though in the present long-wave sets the trailing antenna can be replaced by an auxiliary antenna in such cases, these have a very short range and also, because of the lack of wind, the 70 or 100-watt sets must have a generator driven by an auxiliary motor, which is a great load for the airplane.

Short waves make it possible to meet all these requirements simultaneously. First the power requirement is so low, even at long ranges, that the total weight of the radio installation can be extraordinarily small. Then, the short waves require antenna lengths, which are within the order of magnitude of the present-day dimensions of airplanes, so that good radiating conditions can be obtained even with permanently stretched aerial wires. Finally, a small storage battery can be used for the low primary power, so that the set can be operated even after landing, and particularly after a forced landing.

According to what has been said, one might wonder why short waves were not used in aircraft radio long ago. There have been a large number of objections until the present time. For a long time even the possibility of short-wave reception in airplanes was doubted. It was believed that the ignition noises of the airplane engine, and the vibrations during flight, would make short-wave reception impossible.

But it was primarily the so-called dead zones that made the use of short waves seem to offer very few prospects of success.

For this reason, the propagation of short waves between airplane and ground station was thoroughly investigated by the Deutschen Versuchsanstalt für Luftfahrt.[6] At short and average distances the waves can always be so selected that fading zones do not appear. At greater distances, and especially in transoceanic aircraft communication, waves must be used which show fading zones. The fading of reception is so great that it is impossible to receive in the fading zones with the power available in the aircraft. If direct communication is to be maintained within the same airport station, a change in wavelength during flight cannot be avoided. In long flights or during trips of airships, conditions are more favorable insofar as there is generally not merely one receiving station, but an entire series of stations available for receiving, so that with a given wavelength, it might frequently happen that communication might be broken off with one of these stations while with another ground station, outside the fading zone, communication could be maintained.

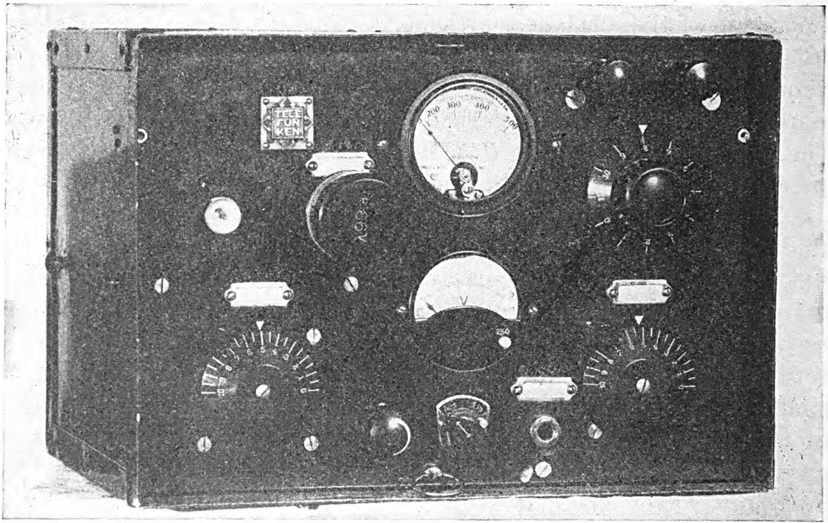

Fig. 12—Airplane short-wave transmitter, DVL Telefunken type.

The construction of short-wave sets for airplanes is still in such a state of flux that a final type cannot yet be described, but in the following we shall briefly describe some examples.

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[6] K. Krüger and H. Plendl, "On the propagation of short waves with low power in the 1000-km region," 124th Report of the Deutschen Versuchsanstalt für Luftfahrt E. V., Berlin-Adlershof, *Jahrbuch der drahtlosen Telegrafie und Telefonie,* 33, No. 3, 1929.

Fig. 12 shows a short-wave transmitter developed by Deutschen Versuchsanstalt für Luftfahrt E. V. in conjunction with Telefunken. Fig. 13 shows its wiring diagram. The first stage is controlled by a quartz crystal. In the second stage the frequency is doubled. In this set the plate potential is controlled in the doubling stage, while the quartz-controlled stage is always oscillating. The second stage has a wavelength range from 30 to 70 m. The pure beat note produced by such a transmitter can be heard above the disturbing noises. An improved form of this transmitter was used by the Deutschen Versuchsanstalt für Luftfahrt E. V. at the time of its first test in the airship *Graf Zeppelin*. This transmitter will be described specially, in the section on airship sets.

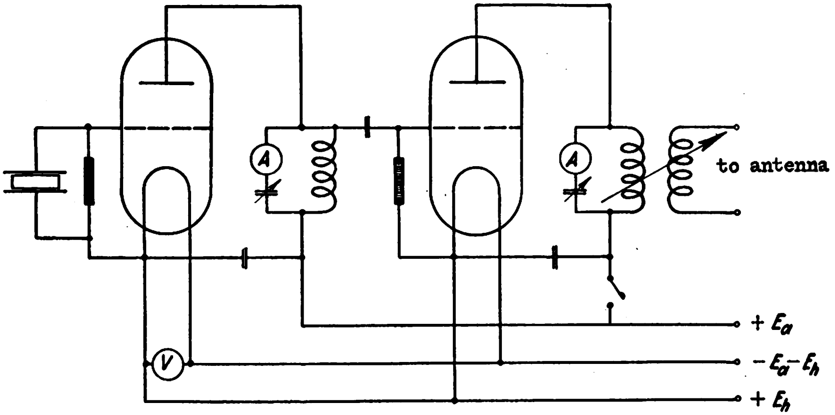

Fig. 13—Schematic wiring diagram of a quartz-controlled transmitter with frequency doubling.

Short-wave reception causes considerably greater difficulties than transmitting in the airplane. These consist, first, in the particularly unpleasant disturbances caused by the ignition spark of the engine, and then especially in the vibrations which cause fluctuations of the frequency of the heterodyne wave, and pitch fluctuations or complete cessation of tone in the receiver.

The first difficulty can be overcome by suitable screening of the ignition wires. The disturbances caused by vibrations cannot be removed even by very stable construction of the receiver. For this reason a quartz-controlled short-wave receiver with long-wave intermediate frequency was developed by the Deutschen Versuchsanstalt für Luftfahrt E. V.[7] In this receiver the heterodyne frequency of the receiver should be stabilized and then the pitch of the receiving tone can be adjusted and readjusted at will, if any change should occur in heterodyne tone due to a temperature change or other influences.

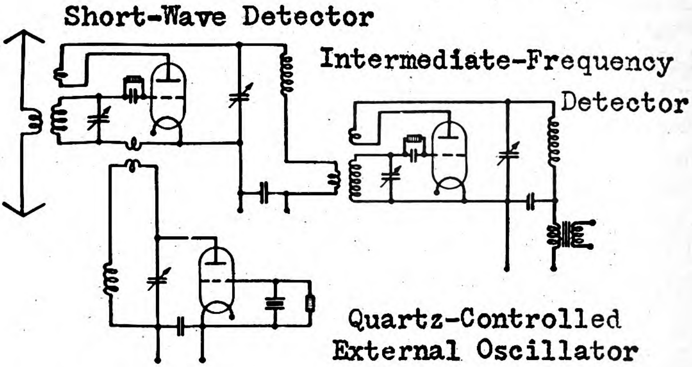
Short-Wave Detector
Intermediate-Frequency Detector
Quartz-Controlled External Oscillator

Fig. 14—Quartz-controlled short-wave receiver with intermediate receiver.

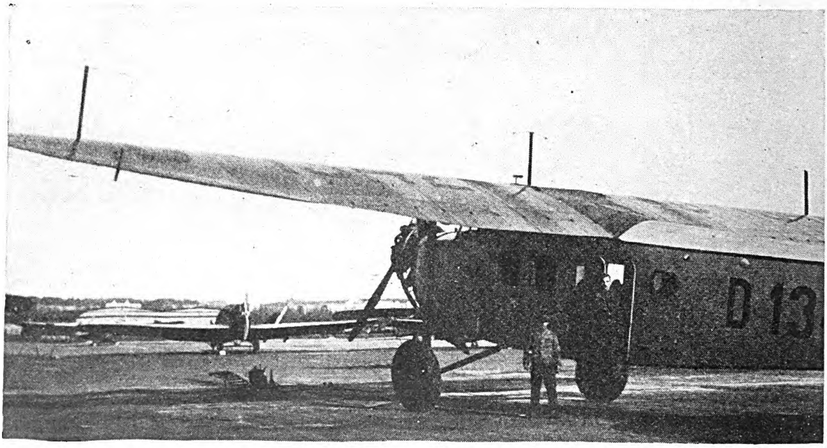

Fig. 15—Transverse dipole on a Möve.

For this purpose, the quartz wavelength of the receiver is so selected in relation to that of the transmitter, as to produce beats of radio frequency. This r-f beat frequency is received by a tube detector.

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[7] P. v. Handel, K. Krüger, and H. Plendl, "Quartz-control of short-wave receivers," 141st Report of the Deutschen Versuchsanstalt für Luftfahrt E. V., Berlin-Adlershof, *Jahrbuch der drahtlosen Telegrafie und Telefonie*, **34**, No. 1, 1929.

If the quartz wave of the receiver varies, the pitch of the receiving tone in the long-wave tube detector can be readjusted without difficulty.

Fig. 14 shows the wiring diagram of such a receiver. It is essential that the first (short-wave) detector should not oscillate and its regenerative coupling should only be used for the reduction of damping.

All difficulties in short-wave reception in airplanes can be removed if, according to the proposal of the Deutschen Versuchsanstalt für Luftfahrt E. V., the short waves are used only for transmitting from the airplane, while long waves are used from the ground to aircraft. The disadvantage resulting from this, namely, the use of a ground transmitter with very high power, is less important if ground stations are available with sufficient power for long distance communication.

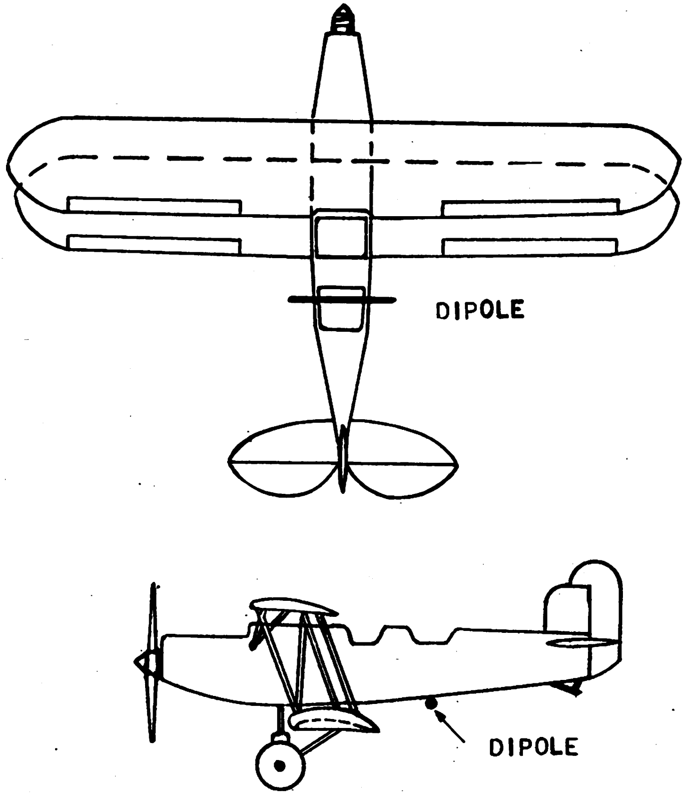

Fig. 16—Position of the dipole on the airplane.

Securely stretched dipoles are most commonly used by Deutschen Versuchsanstalt für Luftfahrt E. V. Fig. 15 shows the installation of such a dipole antenna on a Lufthansa airplane. The antenna as tuned to a fixed transmitting wavelength can be used for receiving on any wavelength as the receiver described above works with an aperiodic antenna. The aerodynamic resistance of this aerial was determined by the Deutschen Versuchsanstalt für Luftfahrt E. V.[8] It is approximately equal to that of a trailing antenna 70 m long.

Up to the present time short waves have been tested in Germany only superficially in the operations of the German Lufthansa. According to past experiences, short waves are particularly suitable for radio communication over very long distances. It will be the problem of the immediate future to perfect short-wave apparatus more and more in order that the reliability of short-wave operation will be equal to that of long waves in aircraft communication at the present time.

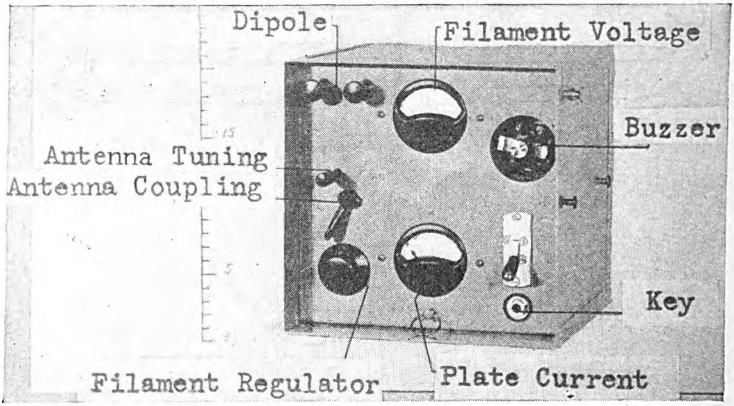

Fig. 17—Telefunken transmitter for 3.7 m.

IV. THE USE OF ULTRA-SHORT WAVES

Ultra-short waves have not been tested up to the present in the German aeronautical radio service, but tests have been made in airplanes[9, 10] equipped with these waves. Thus it was intended to determine experimentally whether the idea of the so-called optical propagation of these waves corresponded to fact, and this question was to be answered by range measurements in relation to the height of the airplane above the earth. The extent to which ultra-short waves can be used in aviation depends on the result of such investigations.

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[8] F. Liebers, "On the resistance of airplane antenna and the reduction in flying power caused by it," 100th Report of the Deutschen Versuchsanstalt für Luftfahrt E. V., Berlin-Adlershof, *Luftfahrtforschung*, **1**, No. 4, 1928.

[9] H. Fassbender and G. Kurlbaum, "The dependence of the range of very short waves on the height of the transmitter above the earth," 125th Report of the Deutschen Versuchsanstalt für Luftfahrt E. V., Berlin-Adlershof, *Jahrbuch der drahtlosen Telegrafie und Telefonie*, **33**, No. 2, 1929.

[10] F. Gerth and W. Scheppmann, "Investigations on the propagation processes of ultra-short waves," *Jahrbuch der drahtlosen Telegrafie und Telefonie*, **33**, No. 1, 1929.

A small transmitter built by the Telefunken G. m. b. H. for a wavelength of 3.7 m was placed in the wooden airplane of the Albatross L 74 type. The maximum height reached by the airplane during the test was 2700 m (9000 ft.). The antenna was a dipole 170 cm long, made of copper tubing which, as seen in Fig. 16, was placed below the fuselage perpendicular to the axis of the airplane. The connection from the transmitter to the dipole consisted of an electric line screened by a metal cylinder. In Fig. 16 we see plainly the small dimensions of the antenna tube as compared with the airplane, so that the airplane could not be hindered by this, as with the normal long-wave trailing antenna.

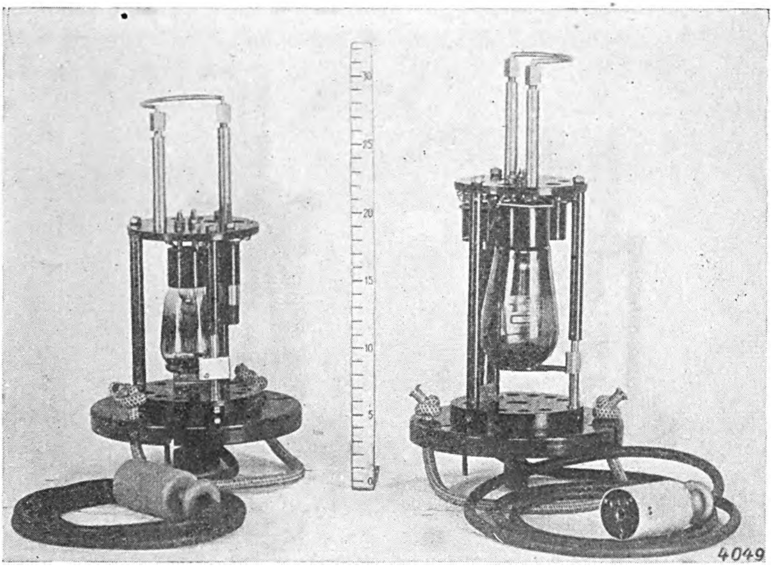

Fig. 18—Lorenz transmitter for ultra-short waves; high-frequency part of transmitter and receiver.

Fig. 17 shows the Telefunken transmitter used in the previous experiments. Fig. 18 shows the high-frequency part of the transmitter and receiver of the Lorenz set. The antenna is the small loop shown above the tube. This part must be so placed in the airplane that the passage of the electromagnetic waves to the receiving station is not hindered by parts of the body of the airplane. Fig. 19 shows the low-frequency part of the set (arrangement for modulation and super-regeneration).

TABLE III
LIMITS OF TELEGRAPHIC RECEPTION WITH $\lambda = 3.7$ M, IN RELATION TO FLYING ALTITUDE (AIRPLANE FLYING AWAY FROM THE RECEIVER)

Flying Altitude (meters)	Limits of Telegraphic Reception (km)
190	36
200	34
220	36
450	54
510	54
680	67
890	65
2620	130

Table III and Fig. 20 tabulate the limits for telegraphic reception in relation to the flying altitude for reception on the ground and in the airplane. The results were obtained with a super-regenerative receiver.[11] According to the idea that with these waves there must be optical vision between transmitter and receiver, reception will be possible as long as the airplane is above the plane tangential to the surface of the earth at the place of reception. If we designate the flying height by h (Fig. 21), the range by a, the radius of the earth by r we get

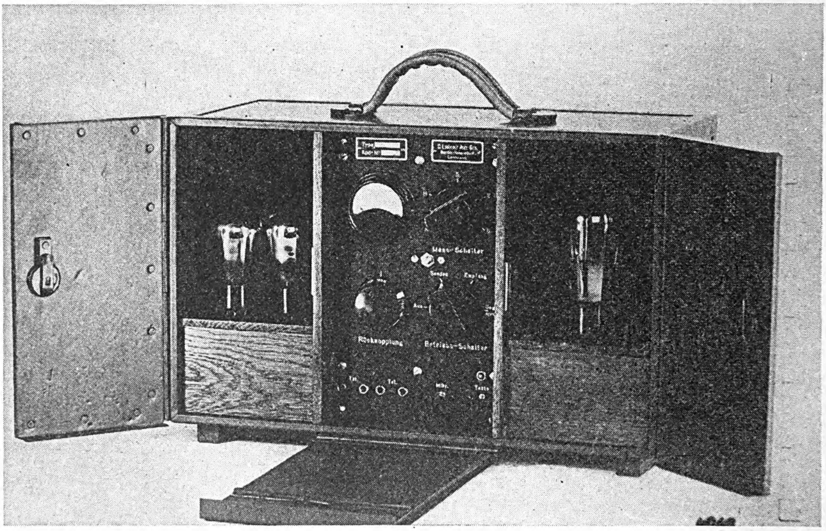

Fig. 19—Lorenz transmitter for ultra-short waves. Low-frequency part.

$$\cos\varphi = \frac{r}{r+h} \text{ and } \varphi = \frac{a}{r}$$

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[11] E. Busse, "Super-réactions," *L'Onde Electrique*, **7**, 217–259, 1928. Extensive literature, *Jahrbuch der drahtlosen Telegrafie und Telefonie*, **33**, 153, **1929.**

For $a \ll r$ we get $a = 2\ rh$. For the radius of the earth $r = 6{,}370$ km; the theoretical range $a_{km} = 3.55 h_m$.

The values thus calculated are drawn as dashed lines in Fig. 20, while the solid curve gives the ranges actually found.

Ultra-short waves seem to be particularly well suited for communication between two airplanes.

A certain competition is to be expected in this field of application of ultra-short waves by the waves between 100 and 150 m. The radiation is extraordinarily great for these waves also. The range can be regarded as definite, because of the small space radiation. Fading is present it is true, but is endurable. Communication with distant ground stations is not possible with these waves, as the waves traveling along the surface of the earth are subject to extraordinarily large attenuation. But nothing prevents the use of these waves for communication between two airplanes. Whether the waves below 10 m or those between 100 and 150 m are the most suitable for this field of application, cannot be decided at the present time, and depends upon the development of receivers for these two wave regions. It must be mentioned, as an essential advantage of the 3-m wave, that spacial damping due to the ground can be disregarded at relatively low flying heights.

V. DIRECTION-FINDING APPARATUS

In order to maintain regular aerial transportation, the pilot at at each instant must have an exact knowledge of his location, or at least of the course being followed. Radio bearings are an important aid to navigation in the airplane.

All radio direction-finding methods can be divided into two large groups: direction-finding by a fixed station and direction-finding with the aid of a radiocompass on the plane. The German air transportation system has a number of ground direction-finding stations which determine the location of an airplane by bearings on request and then radio it to the airplane. The long-wave airplane transmitter used for communication is used as a bearing transmitter. The ground direction-finding stations all have Telefunken goniometers and use the apparatus that was developed a long time ago for German ships. The ground installation consists of a receiver, an 8-tube set with quadruple high-frequency amplification, a detector and triple low-frequency amplification, a loop-antenna which can be rotated, the azimuth of which can be read to 1/2 deg. on the bearing disk, and the auxiliary or sense-finder antenna. The sense-finder is necessary to compensate for the so-called antenna effect of the goniometer, which is due to the fact that in the loop there are produced not only the re-

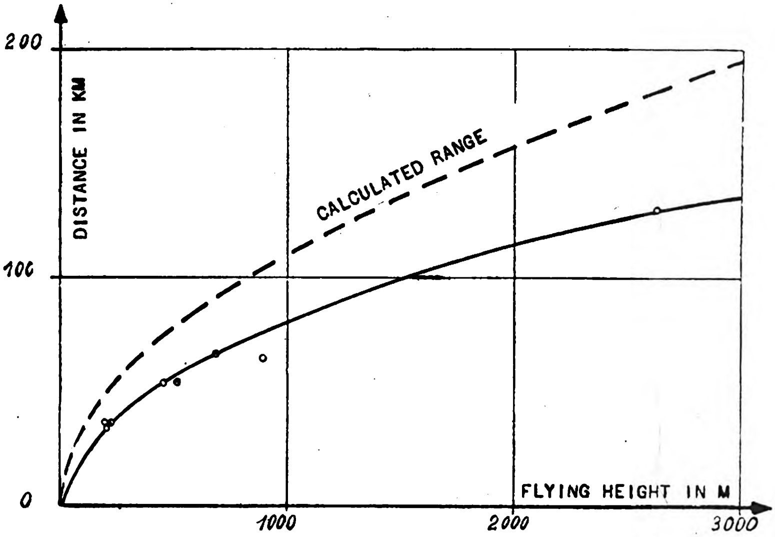

Fig. 20—Limits of telegraphic reception, depending on flying height—flying away.

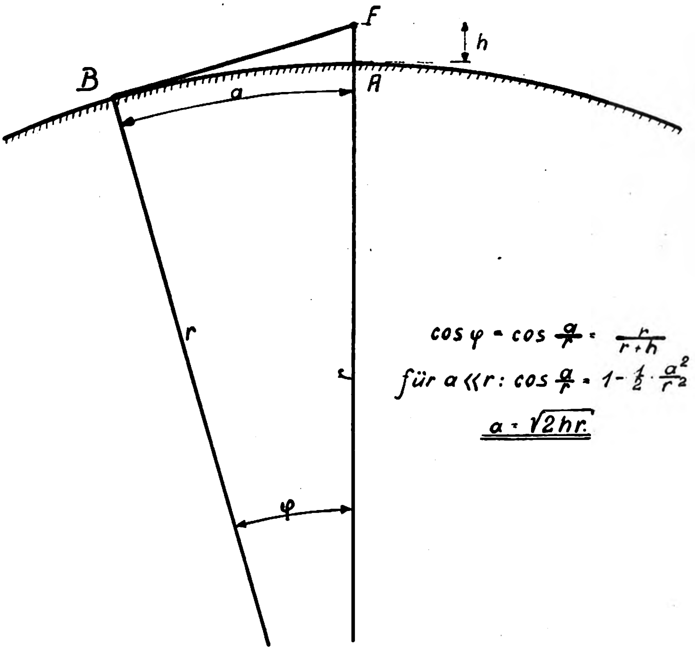
F
h
B
A
a
r
r
φ
$\cos\varphi = \cos\frac{a}{r} = \frac{r}{r+h}$
für $a \ll r$: $\cos\frac{a}{r} = 1 - \frac{1}{2}\cdot\frac{a^2}{r^2}$
$a = \sqrt{2hr}$

Fig. 21—Calculation of theoretical range.

ception alternating potentials due to the magnetic field, but a potential is also produced by the electric field which is present simultaneously. If the loop is turned to determine the direction of minimum reception, the potential caused by the antenna effort remains, and frequently causes considerable indistinctness in the position of the minimum. This error is compensated by the fact that, due to the auxiliary antenna, there is induced in the loop circuit, a potential equal to the antenna effect and with a phase of displacement 180 deg. By the correct adjustment of this auxiliary antenna and the loop, extraordinarily sharp minima are obtained.

As the characteristic curve of the direction-finding shows two minima which are displaced 180 deg., it is necessary to make a lateral determination for the establishment of the airplane location. Here the sense-finder is used, the characteristic curve of which is circular. Superposition of this characteristic curve on that of the loop, with the correct balancing, gives a unilateral curve with only one minimum, from the position of which it can be determined on which side the loop of the airplane is located.

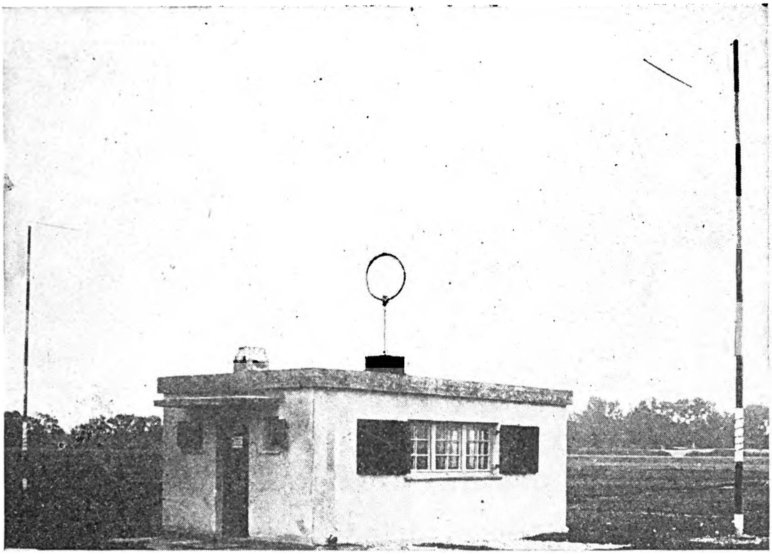

Fig. 22—Munich ground direction-finding station. Outside view.

Direction-finding by a fixed station has the advantage that no additional apparatus is necessary in the airplane, and that the airplane crew is not bothered with difficult plotting problems. It is proving valuable in Germany to an increasing extent. Its disadvantage is that such a direction-finding method requires a certain amount of time so that, particularly in bad weather, if determination of locations are frequently necessary, one airplane must wait for another, and the value of such bearing determinations by fixed stations can be greatly reduced under certain circumstances.

The ground direction-finding stations are placed in small brick houses which are located in such a position on the flying field that the disturbing influences of neighboring aerials are as slight as possible. A view of the ground direction-finding station in Munich is shown in Figs. 22 and 23, in which the rotating loop and the auxiliary antenna are clearly visible.

In direction-finding by means of radiocompasses in the plane the determination of location is made in the airplane. For this purpose a receiver sensitive to direction must be on board. The radiocompass used for a long time in ships and in an improved form in airport ground direction-finding stations, a somewhat lighter type of which was also used in zeppelins, has recently been developed into the airplane compass[12] by the cooperation of Deutschen Versuchsanstalt für Luftfahrt E. V. and Telefunken, and promises to be a necessary navigation instrument for long flights. It has proved its suitability for this purpose in many flights made by Deutschen Versuchsanstalt für Luftfahrt E. V. and Lufthansa.

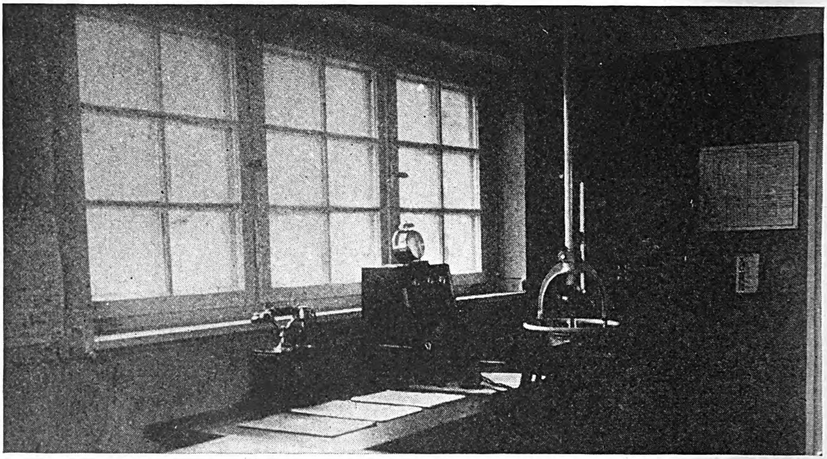

Fig. 23—Munich ground direction-finding station. Interior view.

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[12] M. H. Gloeckner, "The radio compass receiver in the airplane," 126th report of the Deutschen Versuchsanstalt für Luftfahrt E. V., Berlin-Adlershof, *Jahrbuch der drahtlosen Telegrafie und Telefonie*, **33**, No. 4, 1929.

The new airplane radiocompass consists of a light-weight rotating loop, the design of which has been carried out with special consideration

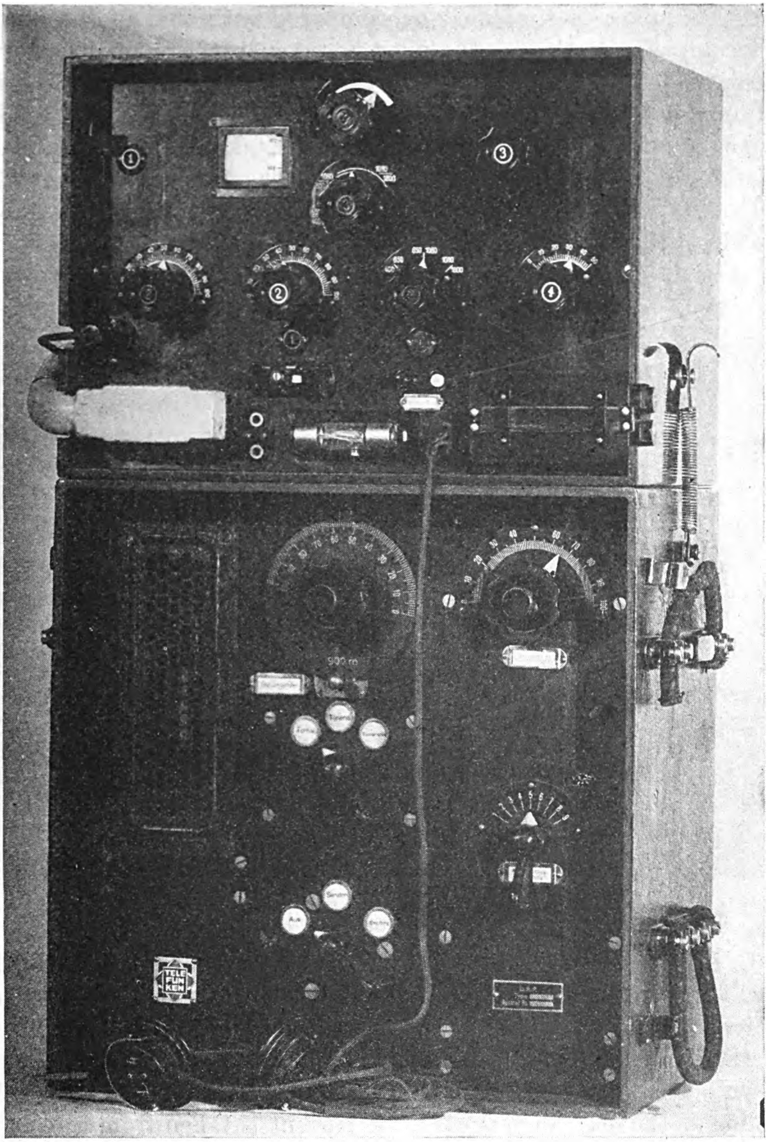

Fig. 24—Airplane transmitter, type 262F, with radiocompass.

of its wind resistance. The radiocompass set, which operates as an ordinary airplane receiver when taking a bearing, has nine tubes and the auxiliary-antenna requirement is taken care of by an antenna fixed on the airplane, whose electromotive forces must be increased by a special tube. The two tuned circuits have single-knob control. The method of sense-finding is simplified as compared with that of the ground radiocompass. As the bearing must be taken by the radio operator on board, an indicator apparatus has been developed whereby the pilot can quickly be informed of his course in a very simple manner. The current supply for the radiocompass set is taken from the propeller generator of the airplane radio set, which has a special current filter for this purpose. Fig. 24 shows such a radiocompass, the dimensions of which are so selected that it can be placed on the transmitter like the receiver of the Telefunken set 262 F.

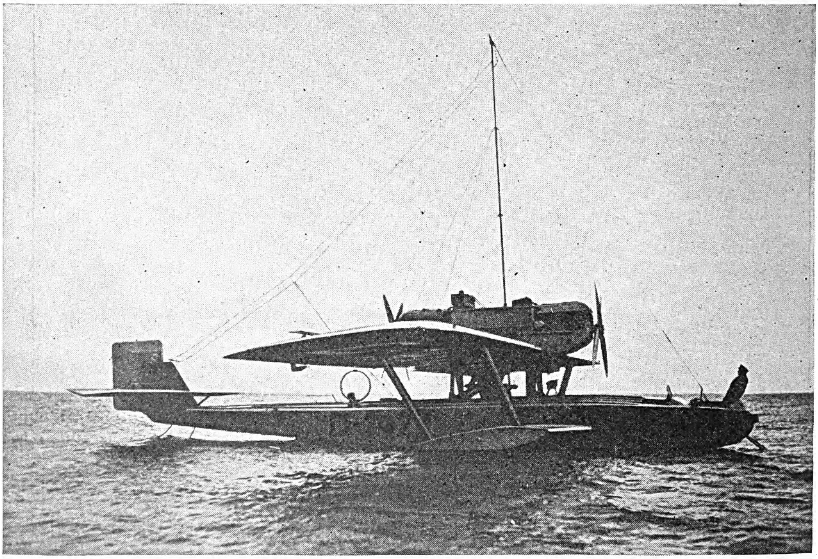

Fig. 25—Dornier Wal with emergency antenna and compass loop.

In Fig. 25 we see the direction-finding loop mounted on a Dornier-Wal. The closeness of the adjacent metal mass causes errors in direction-finding, which are taken into account by drawing calibration curves. On the Wal illustrated, the emergency transmitting mast has been drawn out and serves for maintaining radio communication after landing on the sea. In this case, the current is taken from a benzine-electric emergency transmitting unit.

At the present time the Telefunken radiocompass is the only one that is used to any great extent on the ground or on the airplane as a direction-finder in German aviation. Work is being done on the development of other direction-finding methods. We shall mention here two methods that are possible for direction-finding from the airplane.

It has long been desired to have a direct-indicating direction-finder in aerial navigation for flights to a given point, which makes it possible with only one adjustment to read on the indicator of the instrument in the pilot cabin, whether the airplane is deviating from the direction to a radiobeacon. Experiments along this line have been made by Dieckmann and Hell.[13]

The Lorenz firm in Germany has also made tests on the method of directional transmission. The practicability of this method is being tested at the present time by Deutschen Versuchsanstalt für Luftfahrt E. V. This method, described by Scheller[14] and Kiebetz,[15] has been developed further in America. It has the advantage that a relatively small receiving set on the airplane is sufficient to maintain at all times the direction toward or away from the directional transmitter. It has the disadvantage of requiring a special transmitter.

At this point we also wish to mention short-range direction-finding methods which make it possible for pilots to find the flying field and landing area in weather of low visibility if he has been brought to the vicinity of the airport by one of the long-distance direction-finding methods. Radio offers a large number of methods which have been described, but because of the difficulties of this work none of them has led to a decisive result. We may mention here, as examples, the method of low-frequency and high-frequency "leader" cables, which have been used in marine navigation as well as short waves. Short waves can be focussed, but their range is restricted.

VI. AIRSHIP SETS

The special technical requirements for radio communication with airships are as follows:

1. The radio set must make it possible for the airship to keep in communication with the land and coast stations, as well as ships, that are on the course.
2. The radio set should make it possible for the airship to remain in constant direct communication with the home port, if possible.

The first requirement can be fulfilled only by long waves at the present time.

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[13] R. Hell, "Direct-indicating radio-telegraphy direction finding," *Jahrbuch der drahtlosen Telegrafie und Telefonie*, **33**, No. 4, 1929.

[14] O. Scheller, "Special applications of wireless telegraphy," *Jahrbuch der drahtlosen Telegrafie und Telefonie*, **2**, 341, 1909.

[15] F. Kiebitz, "New experiments on radio-telegraphic directional transmission," *Jahrbuch der drahtlosen Telegrafie und Telefonie*, **15**, 209, 1920.

For propagation over the sea, we have the well-known Austin formula which can be applied to communication between airships and ground:

$$E = \frac{0.12 . \pi . J . h_{\text{eff}}}{d . \lambda} \cdot e^{\frac{-\alpha \cdot d}{\sqrt{\lambda}}}$$

E = receiving field intensity in volts per meter
J = transmitting current in amperes
h_{eff} = effective height of transmitting antenna in km
d = distance between transmitter and receiver in km
λ = wavelength in km
α = attenuation constant (over sea = 0.0015)

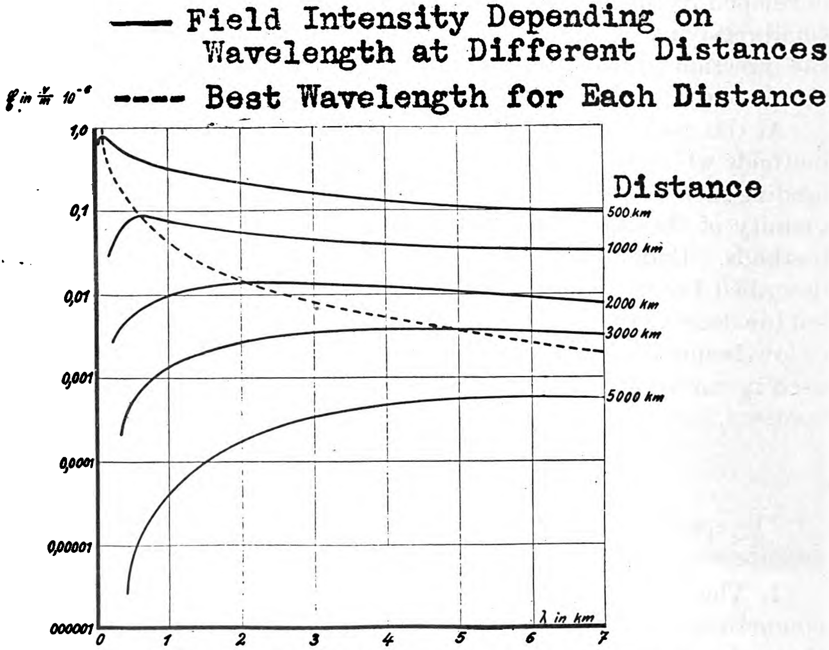

Fig. 26—Calculated field intensities over sea for 1 meterampere, depending on wavelength.

In Fig. 26 there are plotted the field intensities μv per meter according to Austin for different distances in relation to the wavelength, for a current intensity in the transmitting antenna of 1 ampere and for an effective height of transmitting antenna of 1 meter. Fig. 27 shows the receiving field intensities in relation to distance, for different wavelengths, likewise for 1 meterampere. The airship *Graf Zeppelin* has two trailing antennas, one over the other, each 120 m long. The antenna-weights are streamlined. The effective height of the aerials was determined by the Deutschen Versuchsanstalt für Luftfahrt E. V. by field-intensity measurements with the Anders apparatus at the edge of the skip zone. It was about 29 m. This is an especially good value if it is considered that airplane antennas 70 m long, such as are generally used at the present time, have an effective height of only 8.5 m. The antenna current intensity of the long-wave *Zeppelin* set is about 5 amperes. Therefore we can count upon about 145 meteramperes in the transmitter antenna. The field intensities resulting from this are shown graphically in Fig. 28. The curves in Figs. 26 and 28 show that there is a most favorable wavelength for each distance. These values are

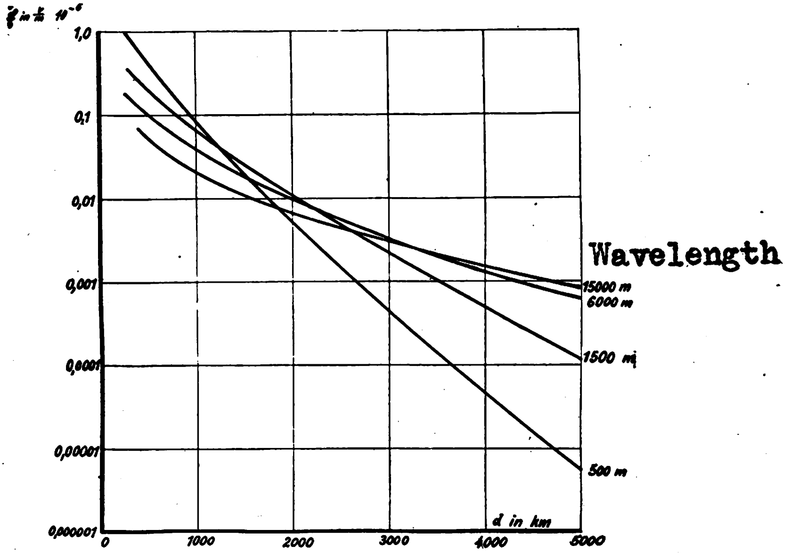

Fig. 27—Calculated field intensities over sea for 1 meterampere, depending on distance.

connected by a dash line in Fig. 26. Therefore, if especially favorable ranges are to be obtained, one must determine the correct operating waves for the different distances.

An electromagnetic field intensity of 3 μv per meter is sufficient for reception on the ground. From Fig. 28 we see that a range greater than 2000 km is not certain in the daytime. Under particularly favorable conditions, and at night, this maximum range can be greatly increased. It must also be remembered that during heavy atmospheric disturbances, very much smaller distances are covered.

The above considerations hold for communication across the sea. The Deutschen Versuchsanstalt für Luftfahrt E. V.[2] has completed an

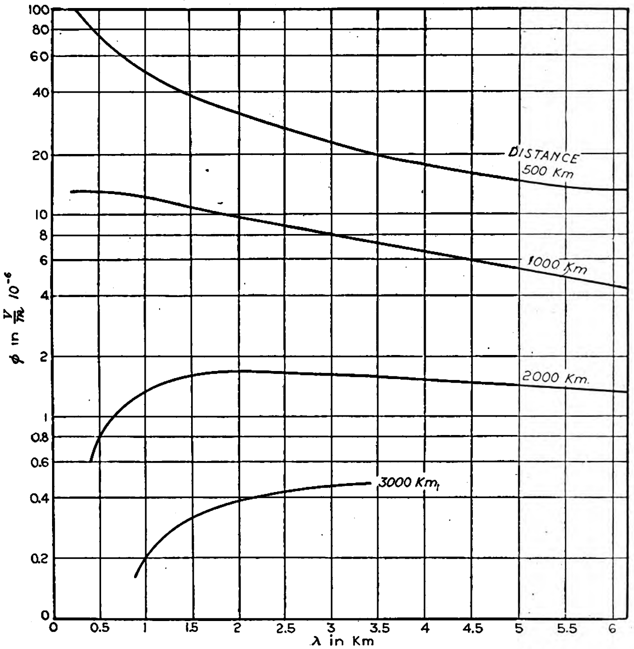

Fig. 28—Calculated field intensities for the Zeppelin long-wave set.

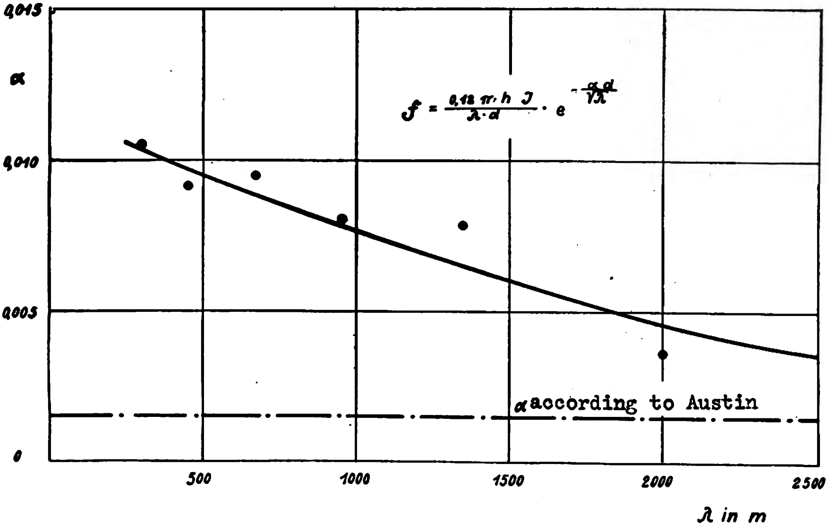

Fig. 29—Attenuation constants α over land.

investigation on long waves and distances up to some hundreds of kilometers over land, which, for communication between airplane and ground, determines the decrease in reception field intensity with distance, for the wavelength range of 300 to 2000 m. The dependence of the attenuation α on the wavelength is shown in Fig. 29. These values gave very small field intensities with the *Zeppelin* antenna. These considerations hold not only for airship radio communication, but also for communication with large airplanes over the long distances, which may be expected.

An increase in the range with long waves would naturally have to be obtained by increasing the effective height of the transmitting antenna and by increasing the antenna current intensity. The use of this method is blocked by the increased load that it would place on the airship. For this reason, the Deutschen Versuchsanstalt für Luftfahrt E. V. made short-wave tests during the first trip of the airship *Graf Zeppelin* to America, the result of which caused the Deutschen Versuchsanstalt für Luftfahrt to install a short-wave set in addition to the long-wave set in the airship.

The radio installation of the airship consists of the following apparatus:

(a) Long-Wave Apparatus.

1. Separately excited Telefunken transmitter type spez 188S, with a power of 140 watts and a continuously adjustable wave range from 500 to 3000 m, which was built basically like the airplane transmitter 262 F, described in detail in an earlier paragraph.

2. Separately excited 70-watt Telefunken transmitter of the airplane long-wave transmitter type 262 F with a wave range from 300 to 1300 m, for use as an emergency transmitter.

Fig. 30, which shows the radio cabin of the LZ 127, also shows these two transmitters on the table at the right.

The current supply for the installation is normally from a generator with self-regulating wind-propeller. In order that it may be possible to transmit even when the propeller is not turning, a second generator is provided, which is fed from a storage battery used for lighting. This battery is normally charged by the wind-driven propeller-generator, and in cases of emergency, by a special benzine unit.

The long-wave set is supplemented by a receiving set which consists of three Telefunken receivers, type E 363S, E 362S, and E 365S with triple-circuit tuning and a wave range from 125 to 25,000 m. Two of these receivers are illustrated in Fig. 31, which shows the other side of the airship cabin.

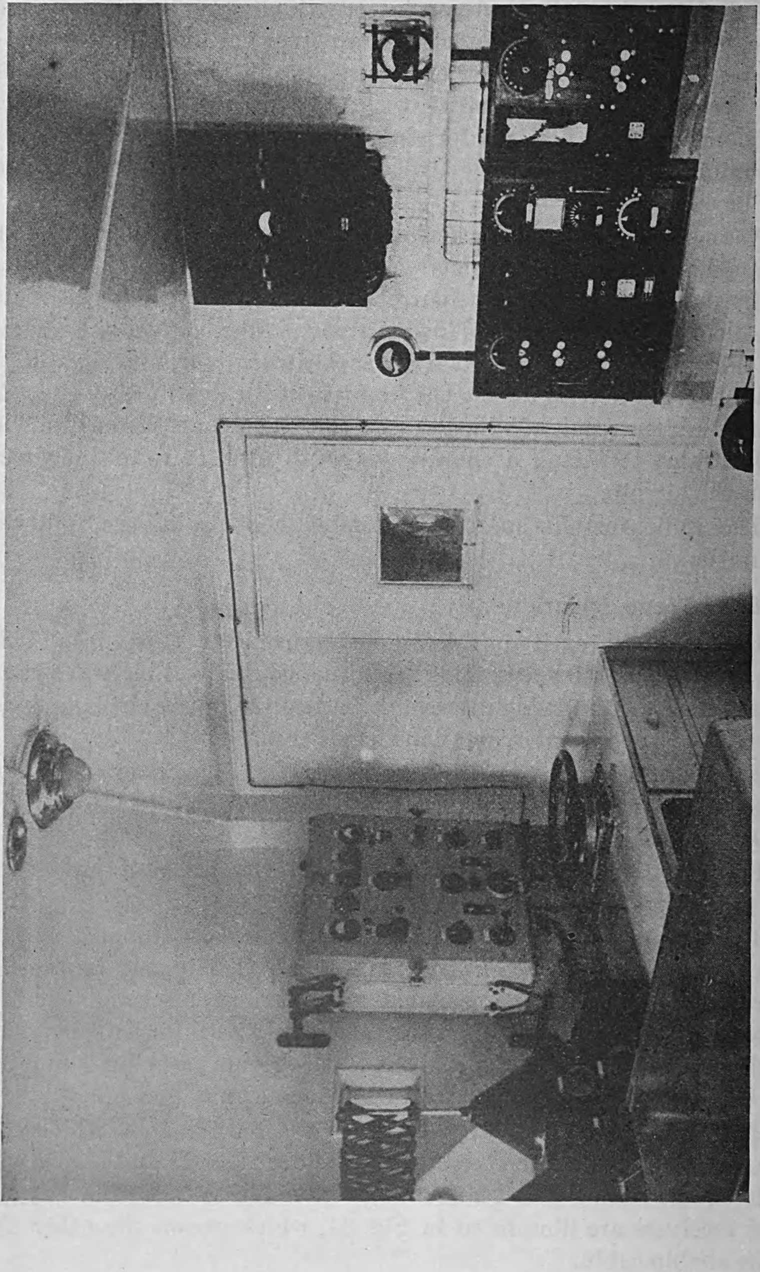

Fig. 30—Radio cabin of the LZ 127, view 1.

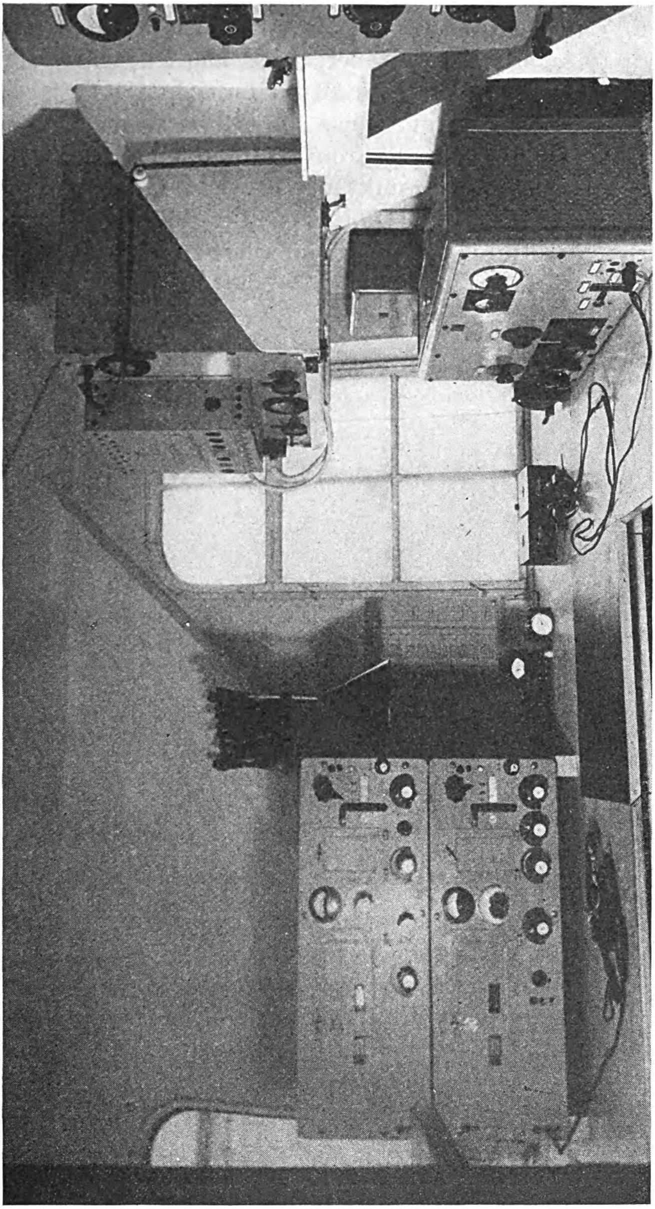

Fig. 31—Radio cabin of the LZ 127, view 2.

(b) Short-Wave Sets.

As stated, the first short-wave transmitter was put in the airship by Deutschen Versuchsanstalt für Luftfahrt E. V. and was in operation on the first trip to America [16] in October, 1928. This transmitter, developed by the Deutschen Versuchsanstalt für Luftfahrt E. V., is shown in Fig. 32. It is quartz-controlled, has one or two stages of frequency-doubling, and has a wave range from 15 to 60 m. During the later trips, the experimental transmitter was replaced by a short-wave Telefunken transmitter. The short-wave transmitter is shown in the top center of Fig. 31. It has an antenna power of about 50 watts, a wave range from 16 to 80 m, and operates just like the two long-wave transmitters with separate excitation. The current supply is taken from the same generators as for the long-wave sets.

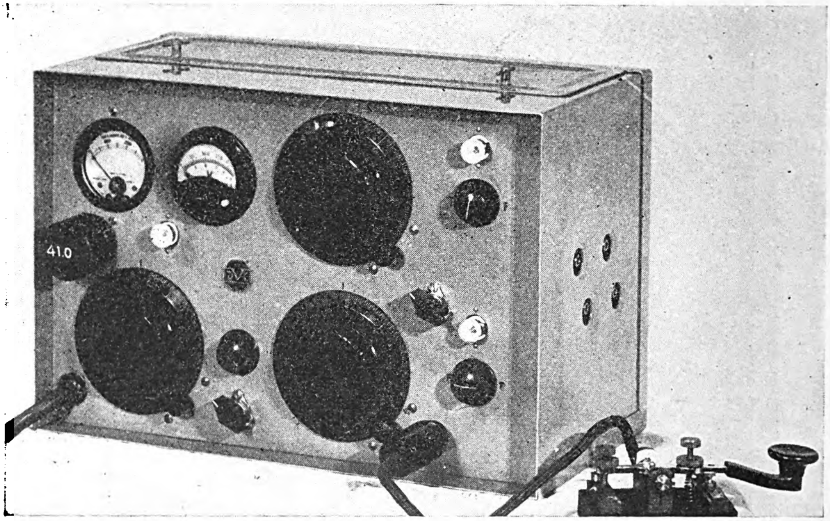

Fig. 32—Experimental airship short-wave transmitter DVL type.

The Deutschen Versuchsanstalt für Luftfahrt E. V. short-wave set operates with the dipole shown in Fig. 33. Although the shape of an airplane generally offers no difficulty in the adjustment of the dipole, the shape of the dirigible is not favorable for this, as can be seen in Fig. 33. The Telefunken short-wave set contains a trailing antenna 15 m long.

The short-wave receiver, which can be seen on the table below the short-wave transmitter in Fig. 31, consists of a 5-tube set containing a regenerative detector in a push-pull connection and two stages of low-frequency amplification. It is equipped with a special oscillator circuit for receiving undamped telegraphy. The wave range of this receiver is 10 to 150 m.

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[16] "Short-wave experiments on the American trip of the airship *Graf Zeppelin,*" 123rd Report of the Deutschen Versuchsanstalt für Luftfahrt E. V., Berlin-Adlershof, *Elektrotechnische Zeitschrift,* 50, 16, 1929.

(c) Radiocompass.

The radiocompass receiver in the airship, which, in its essentials, is similar to the radiocompass previously described, has a wave range from 300 to 4000 m. It can be seen in Fig. 30 on the left beside the door. On the table in front of the radiocompass is the gear for operating the compass loop. The loop is placed below the gondola in the so-called landing buffer.

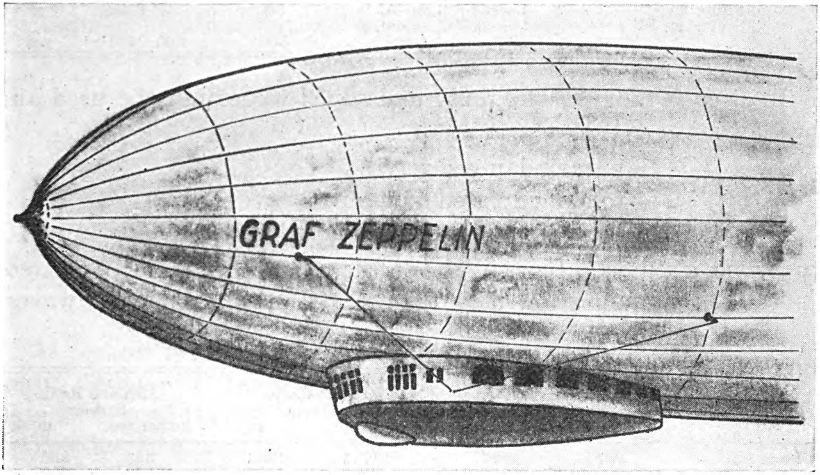

Fig. 33—Dipole on the LZ 127.

VII. OPERATING ORGANIZATION

The operation of the German aircraft radio stations and its organization is invested in a special official board, the "Zentralstelle für Flugsicherung," which is under the Communication Ministry. Ground radio stations have been installed in almost all airports. Table IV gives additional data concerning them. In some airports (see table) direction-finding stations are also installed in addition to these receiving and transmitting sets. Radio operation is controlled by "the operating regulations for international airplane radio together with regulations for the German Airplane Communication Service," agreed upon at the 27th International Aviation Conference in The Hague in September, 1928. Among these regulations, the following are generally important.

TABLE IV

TABULATION OF AIRPORT RADIO STATIONS

No.	Transmitting station	Made by	Antenna power	Wave range	Direction-finder
1	Berlin, FV	Telef. Röhren-Sd.	1500	800–1800	yes
2	Breslau, FV	Lorenz-Röhren-Sd.	400	500–2400	yes
3	Dortmund, FV	Lorenz-Röhren-Sd.	300	500–2400	yes
4	Dresden, FV	Lorenz-Röhren-Sd.	300	500–2400	yes
5	Erfurt, FV	Lorenz-Röhren-Sd.	300	500–2400	yes
6	Frankfurt/M., FV	Lorenz-Röhren-Sd.	300	500–2400	yes
7	Halle-Leipzig	Lorenz-Röhren-Sd.	300	500–2400	—
8	Hamburg, FV	Lorenz-Röhren-Sd.	300	500–2400	yes
9	Hannover, FV	Telef. Röhren-Sd.	1500	800–1800	yes
10	Hof	Telef. Röhren-Sd.	800	800–1800	—
11	Köln	Telef. Röhren-Sd.	800	800–1800	—
12	Königsberg, FV	Telef. Röhren-Sd.	1000	800–1800	yes
13	Lindenberg	Lorenz-Hochfrequ. Masch.-Sd.	5000	800–1900	—
14	München, FV	Lorenz-Hochfrequ. Masch.-Sd.	1000	800–1900	yes
15	Nürnberg, FV	Lorenz-Röhren-Sd.	300	500–2400	yes
16	Stettin, FV	Lorenz-Röhren-Sd.	300	500–2400	—
17	Stolp, FV	Lorenz-Röhren-Sd.	300	500–2400	—
18	Stuttgart, FV	Lorenz-Röhren-Sd. (Röhren =tubes)	300	500–2400	yes

In aircraft radio service, only undamped waves may be used and these are:

pure undamped waves
tone-modulated undamped waves
speech-modulated undamped waves

Under no circumstances will spark (damped) waves be permitted.

Radio communication may take place on the following waves:

TABLE V

	Aviation Weather Radio Service		Flight Radio Service		Airplane Radio Service	
	kc per sec.	m	kc per sec.	m	kc per sec.	m
Long waves	207.5	1444	217.5	1380	323	930
	228	1316	243	1235	333	900
	233	1288	248	1210	345	870
	238	1260	273	1100	—	—
Short waves	2830	106	2804	107	5660	53
	6977	43	5455	55	6593	45.5
	9380	32	7407	40.5	11111	27
	11494	26.1	12000	25	16750	17.9

In the airplane radio service, the waves given in the table will be used under the following conditions:

Calling wave, 333 kc per sec. (900 m) pure undamped, tone and speech-modulated undamped;

Main communication and bearing wave 333 kc per sec. (900) pure undamped, tone and speech-modulated undamped;

Modulation and compass bearing wave, 345 kc per sec. (870 m) pure undamped and tone-modulated undamped.

Modulation and compass bearing wave, 323 kc per sec. (930 m) speech-modulated undamped.

In communication with radio stations of the naval radio service, there will be used the wave 500 kc per sec. (600 m) tone-modulated undamped, and in bearing communication with the same radio stations, the wave 375 kc per sec. (800 m) tone-modulated undamped.

In the service of aviation radiobeacons (radio stations whose transmitting makes it possible for a receiving station to determine its location or a direction in relation to the beacon) there will be used waves in the frequency range 285-315 kc per sec. (1050 to 950 m) pure-undamped and tone-modulated undamped.

Radio communication between airport radio stations is developed fundamentally on telegraphy with undamped, unmodulated waves. Radiotelephony can be used in emergency.

Radio communication between airport and airplane stations is developed fundamentally as follows:

(1) On aircraft whose radio apparatus is not operated by a radio operator: radiotelephony (calling and communication);

(2) On aircraft that have a radio operator: radiotelegraphy on undamped waves as follows:

Airplane station: calling is tone-modulated undamped, communication wave is pure undamped;

Ground station: calling and communication wave, pure undamped.

As regards the use of tone-modulated undamped and speech-modulated undamped transmitting in practice, we wish to refer to the statements made in Section II.

For the purpose of safety in flight, Germany has been divided into fourteen radio communication districts, the radio stations of which have been indicated by "FV" in the table. The airplane will only communicate with the airport radio station in whose radio communication district it is flying, or believes it is flying.

In direction-finding by fixed stations (Frendpeilung) three types of aircraft radio bearings are differentiated:

(a) Target bearing, which is for the purpose of guiding aircraft, by means of an aircraft radiobeacon, from any point to the port at which this beacon is located, by transmitting the deviating course to the objective, once or several times to the airport.

(b) Course bearings, by which the course to any location is communicated after determining the location by two or more direction-finding stations, the course to another location is transmitted to the aircraft.

(c) Location finding in which the location of the aircraft is determined by ground direction-finding stations, and is transmitted to the aircraft by radio.

The radio communication districts of the airport radio stations serve as aircraft radio direction-finding districts.[17] The chief beacon guide station is that airport radio station in whose district the aircraft is flying or in which it believes it is flying.

In direction-finding from radiobeacons, one must differentiate between flight toward a fixed point and the determination of location.

Flight to a fixed object is done very simply with the aid of the radiocompass and requires no special calculations in the airplane. But, on the other hand, it is necessary that a transmitter be in operation at the airport to which the airplane is flying. This can be the airport transmitter, or a special transmitter (radiobeacon for aircraft). If an airplane is to make a flight to a fixed point using its radiocompass, the radio operator sets the loop perpendicular to the longitudinal axis of the airplane after he has first determined the course to be followed by taking bearings on the beacon. If the airplane proceeds on this course, the

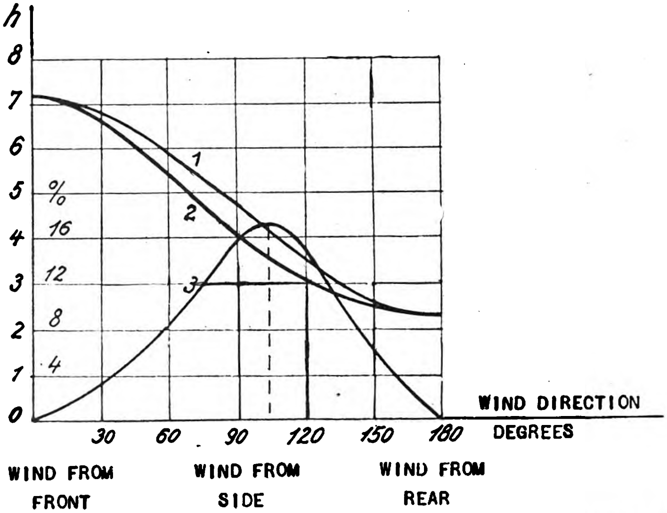

Fig. 34—Time loss in a pure target flight as compared with the great-circle flight.
Curve 1—Flying time without correction angle.
Curve 2—Flying time with correction angle.
Curve 2—Per cent time loss in a flight without correction angle as compared to a flight with correction angle.
Assumptions: flying distance........500 km
airplane speed........140 km per hour
wind speed........... 70 km per hour

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[17] *Nachrichten für Luftfahrer*, 17, 2231, 1928.

operator need only observe that reception is always a minimum in the head piece, and then the airplane is certain to approach its port. Drift, due to wind, keeps it from following the shortest path, but it can be compensated by suitable preliminary setting of the loop. Since the loss in time is very little when the drift is not considered, this compensation is not generally necessary. Fig. 34 shows the loss in time for true target flight as compared with a great-circle flight.

The location of position on an airplane requires bearings from two ground transmitters. The location of the airplane must be calculated from the angles that are measured. The determination of location requires rapid work and considerable calculation on the airplane, so that it is only used when, as in large airplanes for transoceanic flights, there are available other members of the crew experienced in navigation in addition to the radio operators.

Up to the present, direction-finding has been of particular importance in the trips of the zeppelins which are equipped with such radio-compass apparatus.

CONCLUSION

From the above it can be seen that radio engineering plays a very important part in German aviation. Radio sets are required in increasing numbers for transmitting information, for the determination of location and course for safety in flight. Many flights cannot be completed at all without the aid of radio. It can probably be asserted that although one can fly without aircraft radio sets, well-planned air transportation is hardly possible without aircraft radio, and regularity of air transportation cannot be attained without the aid of radio. Aircraft radio engineering serves not merely the flying industry, but it can also be useful in the development of wireless telegraphy in many ways. For example, the aid of the airplane made it possible to take rapid and reliable measurements of attenuation constants, and only with the airplane was it possible to determine the characteristic curve of a short-wave directional antenna, both problems which were taken up by the Deutschen Versuchsanstalt für Luftfahrt E. V. and the results of which are of very general importance for radiotelegraphy.

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THE CONSTANTS OF AIRCRAFT TRAILING ANTENNAS*

BY
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Summary—*This paper constitutes a report on the results of the measurement of the trailing antenna constants on several types of aircraft including the dirigible Los Angeles. A brief description of the apparatus and method is given. Curves for a typical airplane and dirigible are presented and from these curves conclusions are drawn as to the center of radiation and the relation between antenna resistance and the center of radiation. The uniformity of antenna constants of heavier-than-air craft is noted, also the negligible effect of structural differences, speed, and weight on antenna characteristics.*

SINCE the available quantitative information concerning airplane trailing antennas has been limited to constants derived from types of aircraft now obsolete,[1] it seemed desirable to make new determinations with a view to the modernization and extension of the knowledge of this phase of the radio art. Changes in airplane structural design especially in the amount and disposition of conducting elements together with the re-allocation of frequencies, have contributed to the need for accurate antenna information for all types of aircraft.

A survey of the equipment usually employed for the measurement of antenna constants disclosed that none of the existing apparatus could meet the major requirements of space, ruggedness, frequency range, ease of manipulation, and accuracy. Hence, it was necessary to design new measuring equipment, limiting the dimensions to the available space in the smallest of the types of aircraft to be measured.

In its final form the driver, two views of which are shown in Fig. 1, took the form of a 7 1/2-watt Hartley oscillator. The antenna inductance, coupling, resistance, and indicating devices are contained in the unit shown in Fig. 2, which is similar in size to and intended to be clamped on the driver box.

A dynamotor and battery supply power to the driver in the usual case, but in the smaller types of aircraft where space is at a premium a wind or engine-driven generator is employed.

The switching, metering, and regulation of the power supply

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* Dewey decimal classification: R525. Original manuscript received by the Institute, August 21, 1929; revised manuscript received, September 19, 1929.

[1] J. M. Cork, "Airplane antenna constants," **15**, Scientific Papers, Bureau of Standards, p. 199.

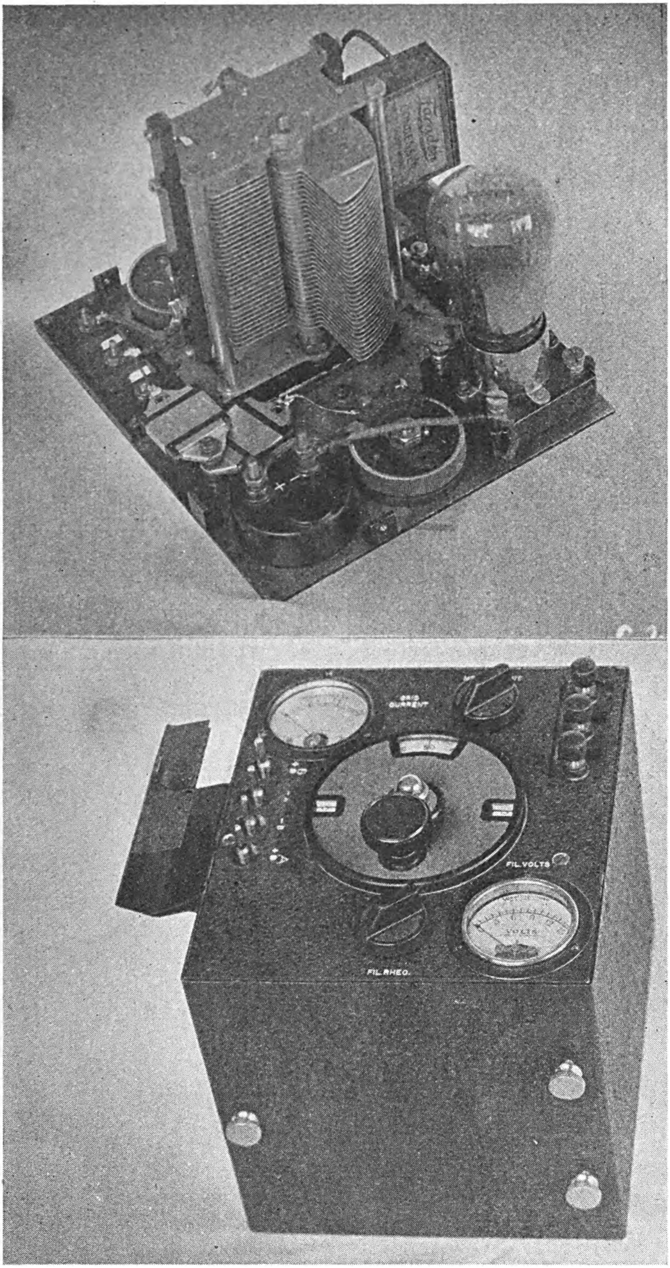
GRID CURRENT
FIL. VOLTS
VOLTS
FIL. RHEO.

Fig. 1—The driver unit, a 7 1/2-watt Hartley oscillator.

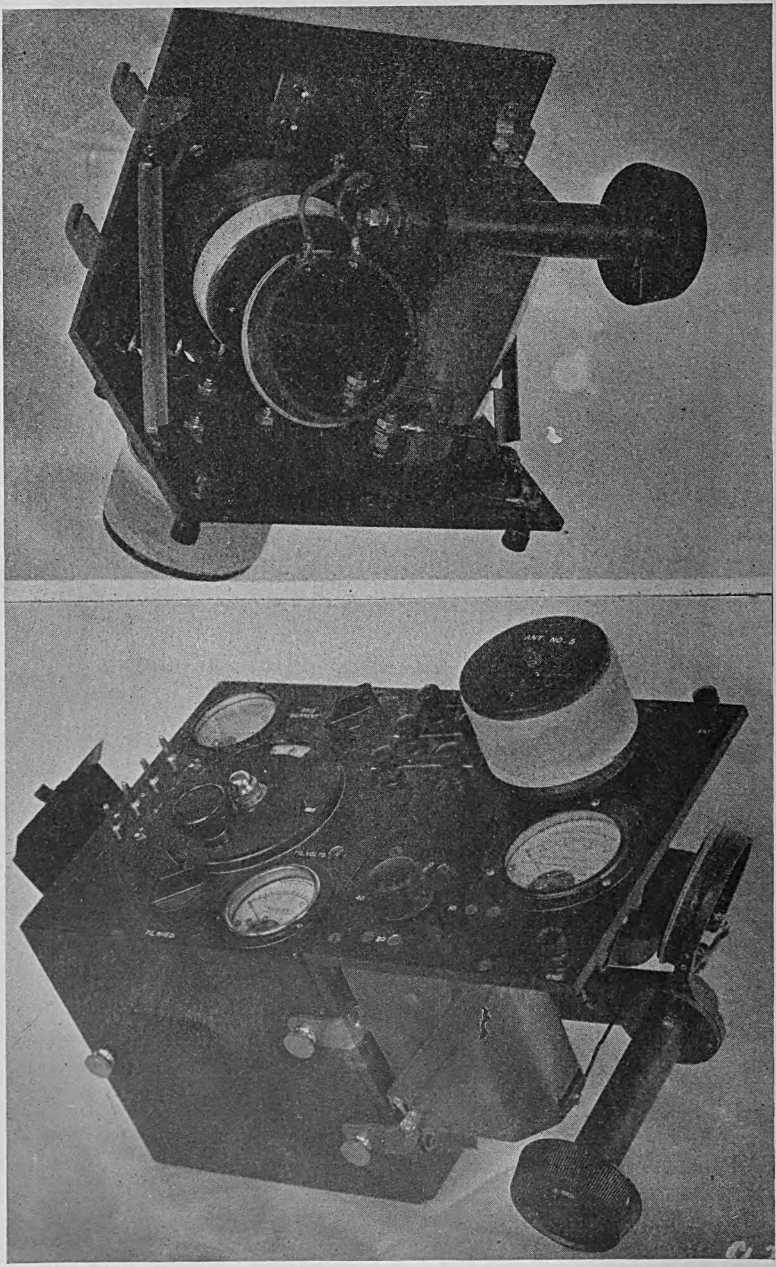

Fig. 2—The antenna unit, with coupler resistors, meter, and jacks for plug-in coils.

is centralized in a small control box adapted to take and distribute the current from any of the several sources.

The entire equipment including wind-driven generator weighs 40 lbs. installed. Control box dimensions are 7 in. x 6 in. x 5 in., while the combined dimensions of the driver and antenna units are 13 in. x 8 in. x 8 in.

By reference to the schematic diagram, Fig. 3, the simplicity of the electrical circuits will be noted. The driver is variable over the range 160 to 4000 kc through the medium of a 0.0015-μf condenser and three plug-in coils. Plug-in coils provide antenna loading inductance variation between 5 and 2500 μh. Considerable care was taken in the design and location of the coils and terminals with the view to eliminating as much of the parasitic capacity, resistance, and coupling as possible.

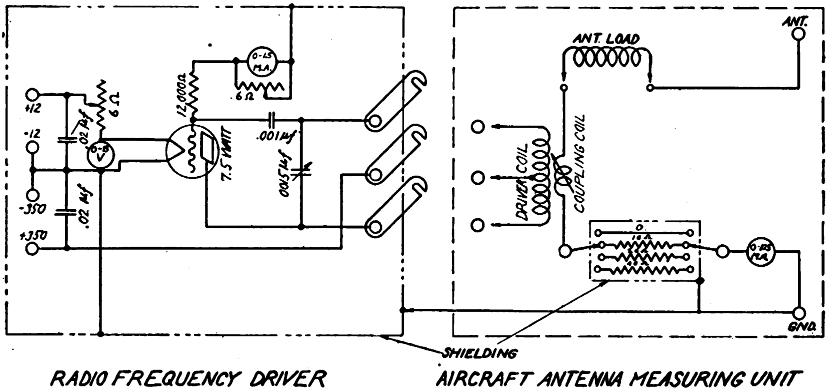

Fig. 3—Schematic diagram of the driver and antenna units.

Tap and dead end losses are avoided through the use of plug-in coils, both in the driver and antenna circuits. These coils, calibrated in the circuit for inductance and resistance over the frequency range in which they are to be used, have proved well suited to the work. Where a minute search of a frequency range is required a variometer is employed. In this case the inductance and resistance calibration are made with the complete instrument set up in operating position.

A novel arrangement of the loading and coupling coils gives a wide variation of coupling, from maximum to zero, simplifies operation, and permits a single coupling coil to be used for the entire frequency range.

Resistance measurements are made by the variation method, three values of non-inductive resistance being successively inserted in the circuit by a switching mechanism controlled from the panel. These resistors are so chosen that in nearly all cases at least one of the inserted values is approximately the same as the combined antenna and circuit resistance to be measured.

On each of the types of aircraft it is the practice to measure the characteristics of the ten lengths of antenna between 50 and 500 ft. in increments of 50 ft. Each length is loaded over the range 5 to 2500 μh, and the resistance and frequency measured for each loading. Flight operations are restricted to the notation of dial and meter readings, all computing being done on the ground. Such an arrangement requires a minimum of flight work—2 hours in the ordinary case where about 250 readings are taken—at no sacrifice in accuracy.

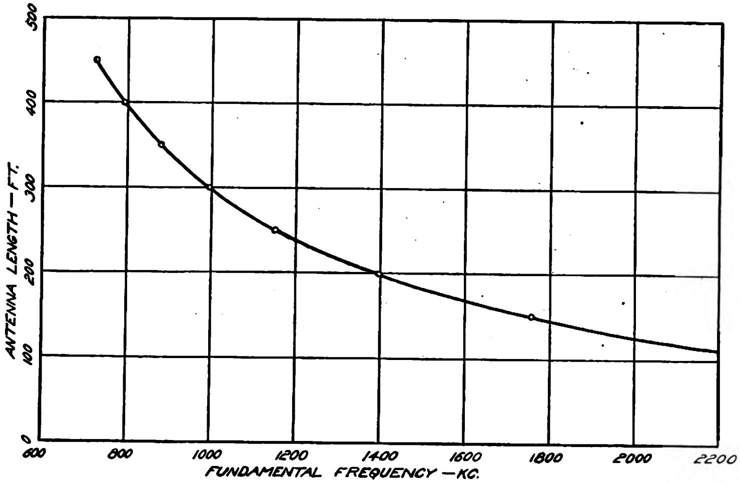

Fig. 4—Fundamental frequency. Ford trimotor all-metal airplane.

Repeated measurements made with the instruments described above agree very closely even when flight conditions vary between a flat calm and strong gusts.

Twelve types of aircraft, varying in size from the single seater pursuit plane to the largest of the patrol flying boats and the dirigible *Los Angeles*, have been covered by the investigation to date. The resultant curves form a voluminous collection not suitable for publication. Instead, a condensed series of curves for the Ford trimotor all-metal ship are shown herein as representative of heavier-than-air ships, while a few of the unique *Los Angeles* curves are given for comparative purposes.

The data obtained from the measurements on the Ford plane and the *Los Angeles* are presented in the form of five series of curves. Two of these are conventional; that is, fundamental frequency vs. antenna length (Figs. 4 and 5), and resistance (Figs. 6 and 7). Figs. 8 and 9 show the loading inductance plotted against frequency of resonance. This curve varies from usual practice though it is similar in scope to the equivalent capacity curve ordinarily given for an antenna. Loading curves have been chosen, however, as more nearly representative of true conditions and as being more informative to

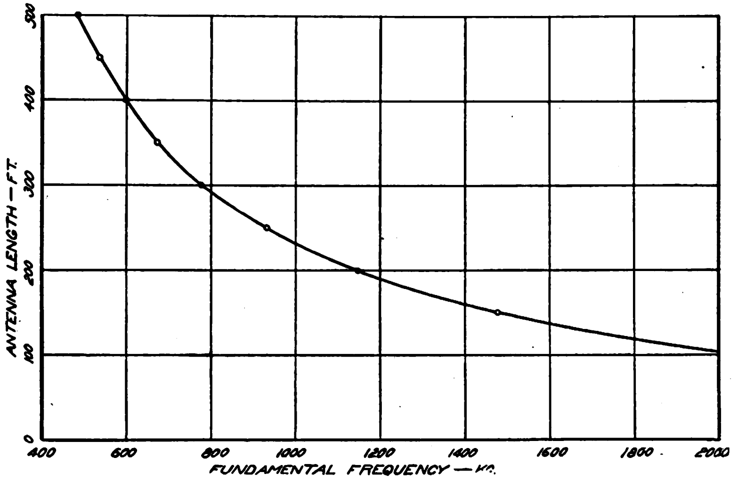

Fig. 5—Fundamental frequency. Rigid dirigible *Los Angeles*.

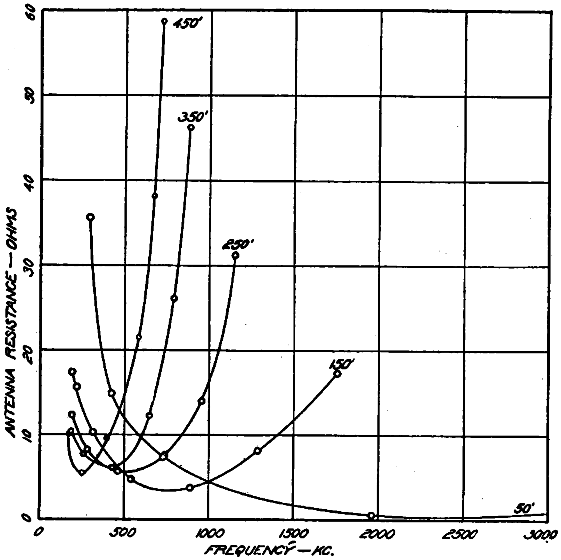

Fig. 6—Antenna resistance. Ford trimotor all-metal airplane.

the designing engineer. The "equivalent capacity" of an antenna is, at best, a nebulous quantity which must be calculated or measured by substitution. At and near the natural period the substituted capacities become large, are not readily obtainable in a form suitable for field use, and are cumbersome and of doubtful accuracy in the laboratory. A substantial variation in the substituted capacity makes little change in the resonance indication due to the low L/C ratio. On the other hand, the inductive load necessary to resonate the antenna at a certain frequency is a finite quantity susceptible to accurate measurement and permanent calibration. In the form of plug-in coils such loading inductances are rugged, compact, and lend themselves readily to field use.

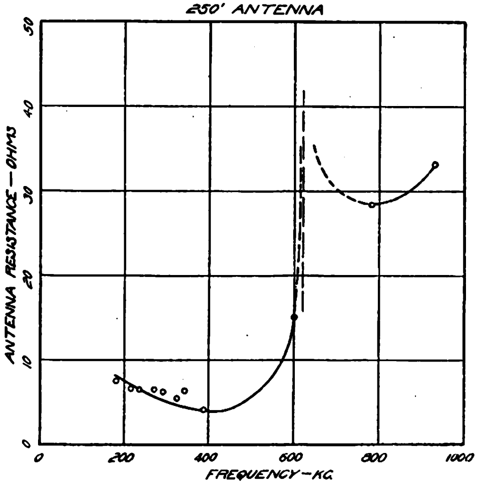

Fig. 7—Antenna resistance. Rigid dirigible *Los Angeles*.

Equivalent and effective capacities and effective inductance may be derived from the loading curves.

In the curves of Figs. 10 and 11 the length of the antenna is plotted against the ratio between one-quarter of the natural unloaded wavelength as measured and the antenna length. Where the ratio is unity the center of radiation is considered to be at the junction between the antenna and the aircraft structure. A larger ratio indicates that the center of radiation[2] lies within the aircraft structure while a ratio less than one points to a current loop somewhere out on the antenna wire. The values which this ratio may assume are proportional roughly to the physical size of the aircraft and may be considered as a good index of electrical size. The table given below shows the length of the trailing antenna at which unity ratio is found on each of several types of heavier-than-air craft. This table forms a convenient means for comparing the "electrical size" or "effective counterpoise" of aircraft.

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[2] For the purposes of this paper the center of radiation is defined as the point in the quarter wavelength antenna-counterpoise system where there is a current loop (voltage node).

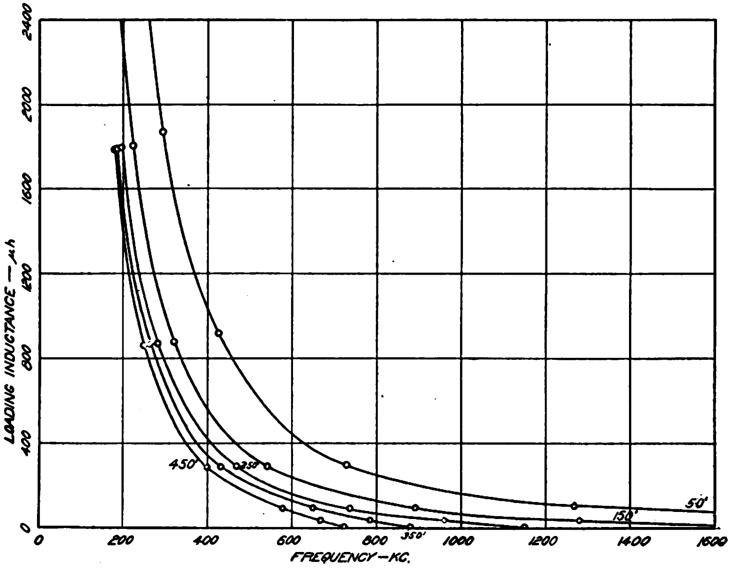

Fig. 8—Inductance loading. Ford trimotor all-metal airplane.

Type	Antenna Length for Unity Ratio $\lambda/4l$
Naval Aircraft Factory H-16 Flying Boat	173 ft.
Loening Amphibian	12∞ "
Sikorsky Amphibian	120 "
Ford Transport	113 "
Vought Corsair	102 "
Curtiss Two Place Fighter	97.5"
Vought Observation (Whirlwind)	93 "

It will be noted that the antenna resistances at the natural period as shown in the curves of Figs. 6 and 7 vary between wide limits for different lengths of antenna on the same aircraft. The values for similar lengths of antenna for different aircraft also disagree. These variations are due to the locations of the center of radiation. In general, where the current loop is at the junction between the antenna and the airplane structure, the resistance at the fundamental frequency will approximate 12 ohms. With the current loop out on the antenna the resistances will be higher and vice versa.

There is a definite relation between the antenna resistance and the length of the trailing wire. If the length of the antenna divided by its natural wavelength be plotted against antenna resistance, both on a logarithmic scale, the result will be a series of points along a straight line as shown in the curve of Fig. 12.[3] From the slope of the line an empirical formula may be developed to show resistance in terms of frequency and antenna length.

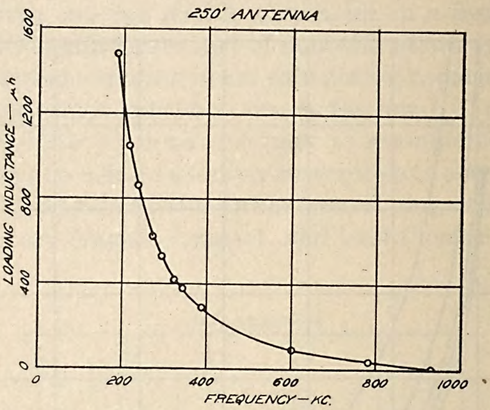

Fig. 9—Inductance loading. Rigid dirigible *Los Angeles.*

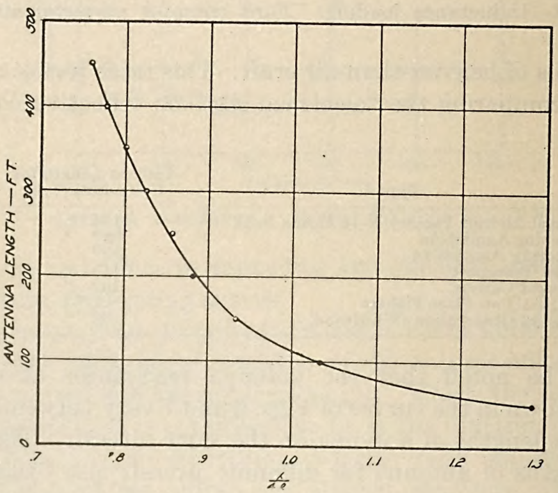

Fig. 10—Center of radiation. Ford trimotor all-metal airplane.

A striking feature of the measurements made on heavier-than-air craft has been the uniformity of the results and the regularity of the curves drawn from the data. Resistance curves compiled from measurements made on the rigid dirigible *USS Los Angeles* bear small resemblance, however, to corresponding airplane curves. Several irregularities are present, and it is believed that the closed loops formed

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[3] For convenience in plotting the value 10 l/λ has been taken instead of l/λ.

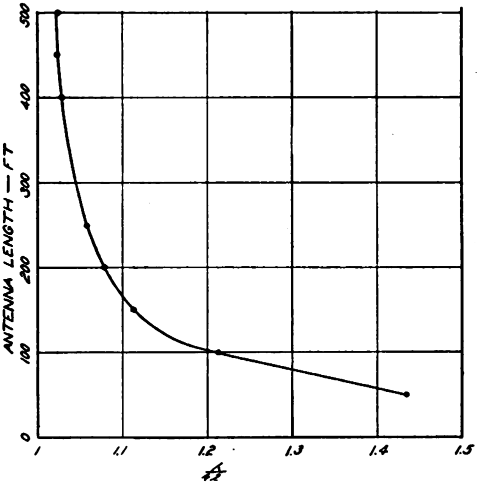

Fig. 11—Center of radiation. Rigid dirigible *Los Angeles.*

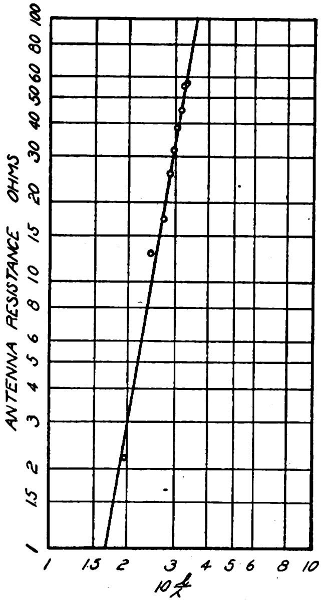

Fig. 12—The relation between the center of radiation and antenna resistance. Ford trimotor all-metal airplane.

by the metal framework are the absorbing mediums responsible for the unique results.

It will be noted also that the curve of Fig. 11 of the *USS Los Angeles* (length vs. ratio length to natural period) is never unity even when the curve is extrapolated to infinity. This is taken to mean that the center of radiation for a structure of such size and type is always within the hull.

An interesting fact developed from this study is the negligible effect of structural differences on antenna constants. The three structural types—all metal, metal frame fabric covered, and composite wood and wire frame fabric covered are interchangeable (when properly bonded) insofar as major effect on antenna constants is concerned. Since the fuselage and wings form a counterpoise of similar proportions, regardless of type, it can be readily seen that the substitution of metal plates for a metallic network should make but little change in counterpoise characteristics. A common ground analogy is met with in large high-voltage condensers where the pipe rack type have nearly the same capacity as plate condensers of the same dimensions.

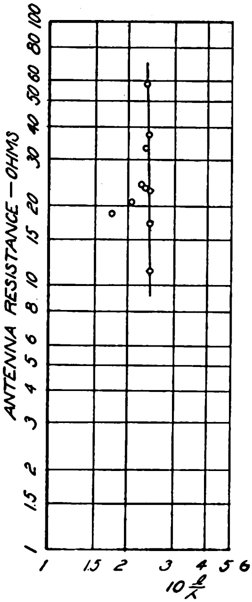

Fig. 13—The relation between the center of radiation and antenna resistance. Rigid dirigible *Los Angeles*.

Some data were obtained on the effect of speed and weight changes on antenna constants. It is well known that changes in the shape and position of the antenna resulting from speed and weight variation produce marked directional effects due to shifts in polarization. Whether there is a corresponding change in antenna characteristics has been a controversial point. In a large plane having a wide speed range a measurement was taken at a speed of forty knots per hour followed immediately by another at 120 knots. A second series of measurements was made with various weights attached to the antenna.

With the longer antenna lengths where shifts in polarization are comparatively large, no measurable change in antenna constants was found under any conditions of speed and weight. A noticeable shift in the resonant point and resistance became evident at the 200-ft. length and was well defined below 100 ft. Apparently the shift in position of that portion of the antenna nearest the plane, say the first 20 or 30 ft., is the cause of the change of characteristics. Such a shift is greater in its percentage change at the shorter antenna lengths but is present without doubt at all lengths of antenna. A case in point was the attempt to measure the characteristics of a Loening amphibian on a very gusty day. Determinations were impossible on the 50-ft. length, but little difficulty was met with in the investigation of the longer antennas.

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ON THE DAYLIGHT TRANSMISSION CHARACTERISTICS OF HORIZONTALLY AND VERTICALLY POLARIZED WAVES FROM AIRPLANES*

By
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Summary—*The investigation described below was carried out for the purpose of comparing the transmission characteristics from an airplane of horizontally and vertically polarized waves in daytime. A frequency of about 6 megacycles was arbitrarily chosen. In view of the fact that modern practice is tending towards the elimination of the trailing wire antenna, an antenna producing vertically polarized waves must of necessity be of small dimensions. The transmissions were therefore compared using a doublet antenna, each arm of which stretched from wing tip to tail for the horizontally polarized wave, and a rigid antenna six feet high for the vertically polarized wave. It was found that the sky ray began to be appreciable with the horizontally polarized wave at a distance of 20 miles, while on the vertically polarized wave it became important at a distance of 50 miles. The signal from the sky ray for distances of the order of 150 miles was always stronger with horizontal polarization than with vertical. The result with the direct ray was different. The signal from the vertically polarized wave was stronger than that from the horizontally polarized wave over highly conducting ground, while the reverse was the case over badly conducting ground. For this reason, the signal from the vertically polarized wave sometimes became very weak at a distance of 40 to 50 miles, before the sky ray was able to arrive with sufficient strength. This occurred particularly on badly conducting ground and when the airplane was flying low. From the horizontally polarized wave, strong signals were received in all conditions continuously for distances up to 600 miles, and there did not appear to be any tendency for the signal to decrease in intensity as this distance was approached.*

A very effective ground antenna was investigated, which possesses the property, when used horizontally, that the signal received is little affected by the height of the antenna above the ground. This permits the use of a very low antenna.

The effect of the conductivity of the ground was considered and, in the appendix, a method is described by which it is possible to deduce the transmission characteristics for flights at any altitude. when those at given altitude are known. The results of the experiments are qualitatively explained on the basis of the present knowledge of transmission phenomena.

WITHIN recent years much data have been accumulated regarding the propagation of electromagnetic waves. Most of the data have been obtained by investigating the variations of the signals received at a given distance from one ground station to another. Where the distance between the stations has been a variable, the results were obtained largely by statistical means from a number of

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* Dewey decimal classification: R115. Original manuscript received by the Institute, November 5, 1929.

ground stations. Such a method is not very satisfactory, for the variations in the ground conditions at the various stations may cause considerable differences in the results, and these differences may often be wrongly attributed to the effect of the distance from the transmitting station. An ideal way of obtaining data giving the variation of signals with distance over land is to employ an airplane as one of the stations. Significant distances can thus be covered in comparatively short time, and the antenna at both the receiving and the transmitting stations is exactly the same at all distances. The data obtained by two methods are not quite the same. The sky ray from a ground station can be expected to be identical in nature with that from an airplane, but the attenuation of the direct ray is very different in the two cases.

In order to obtain as useful data as possible, it was considered desirable to choose a wavelength such that the sky ray could be expected to appear before the direct ray fell to an undetectable value. Then from the results available, the effect of varying the wavelength above or below that employed might be predicted with a fair degree of certainty. An airplane is a comparatively small structure, and it is impossible to install on it a rigid antenna which would be efficient at long wavelengths. On this score it is desirable to use short waves. On the other hand, the sky ray on very short waves does not make an appearance for some considerable distance, while the direct ray is much more rapidly attenuated on short waves than on long waves. A compromise had to be reached, and a wavelength in the neighborhood of 50 meters was finally chosen for the investigation. A longer wavelength might have been chosen if a trailing wire antenna had been employed on the airplane, but, as modern practice seems to be tending towards the elimination of this type of antenna, it was not considered desirable to employ it. If a trailing wire antenna is not used, the antenna for vertically polarized waves is necessarily of very small dimensions. In this investigation, the properties of the horizontally and vertically polarized waves were compared, not with similar antennas on the airplane, but with such antennas as could be reasonably employed in practice. The vertical antenna was a small self-supporting mast 6 ft. high above the fuselage, while the horizontal antenna was a doublet, each arm of which stretched from wing tip to tail of a Fokker Super-Universal monoplane having a wooden wing and a fuselage framework of metal. The general dimensions are shown in Fig. 1. All the results were obtained during the daytime.

Among the data already published, which have a bearing on this investigation, may be mentioned the work of R. A. Heising, J. C.

Schelleng, and G. C. Southworth,[1] between ground stations at various wavelengths. Their results differ from those given below essentially in the distance at which they found the sky ray to become greater than the direct ray. This distance in daytime they found to be of the order of 100 miles, while from an airplane it was found to be 50 miles. The discrepancy is the more astonishing as the direct ray is likely to be more attenuated between ground to ground stations than between ground to airplane.

Not much data have been published for transmissions from airplanes. K. Kruger and H. Plendl[2] made rough audibility tests on low power up to 1000-km distance using a horizontal antenna on the airplane. Their results differ from those given below only in their statement that the altitude of the airplane had no effect on the signal strength. This result, though confirmed by F. Eisner, H. Fassbender, and G. Kurlbaum[3] on long waves, is contradicted by D. K. Martin.[4] On very short waves of 3.7 m, H. Fassbender and G. Kurlbaum obtained results on local flights. The difficulties of measurement at such wavelengths probably account for their failure to obtain agreement between theory and experiment.

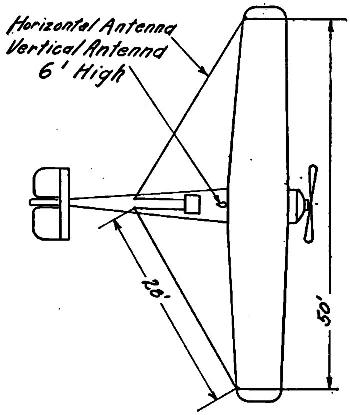

Fig. 1—Antenna arrangement on test plane.

APPARATUS AND METHOD

The vibrations and other physical disturbances associated with airplanes made it advisable to carry out the measurements of the signal strength on the ground and transmit from the airplane. The transmitter used was the same for both the vertical and the horizontal antennas. The antennas were excited at the smallest level of power input, as found by experiment, which would produce a conveniently measurable field intensity at the receiving point, throughout the zones of distance at which successively the ground ray predominated, the ground and sky rays were comparable, and the sky ray predominated. The use of larger currents in the transmitting antennas would not alter the relative extent of these zones. The current at the center of the horizontal doublet antenna was 0.3 ampere corresponding to a power dissipation of approximately three watts, while in the vertical antenna the current was of the order of 0.5 ampere, throughout the experiments. This current was tone-modulated at approximately 200 cycles for convenience in aurally identifying the transmitted signal.

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[1] PROC. I. R. E., 14, 613–647; October, 1926.

[2] "Propagation of low power short waves in the 1000-kilometer range," PROC. I. R. E., 17, 1296–1315; August, 1929.

[3] *Zeits. f. Hochfreq.*, 31, 109–116; 1928.

[4] "Laying the foundation for aircraft communication," *Bell Laboratories Record*, 7, 315–318, 1929.

Two methods are available for measuring signal or field intensities at 6 megacycles. A radio receiver and output instrument may be used as a comparator of received voltage with known voltages from a local adjustable source; or a receiver of suitable design provided with an adjustable sensitivity control and an output meter may be calibrated in voltage sensitivity as a function of the sensitivity control and used directly as a measuring device. Both methods derive their ultimate accuracy from the original calibration of a local signal source. The latter involves two calibrations and is subject further to accidental variations. It is experimentally cumbersome on this account, but this disadvantage is offset by the highly important consideration that during an experimental run readings may be taken much more rapidly than with the comparator scheme. Since the present measurements were made under simultaneous conditions of fading and rapidly changing distance, preliminary experiments quickly emphasized the desirability of the direct-measurement method. The simplicity of the apparatus was another desirable feature.

A tuned carrier-frequency amplifier was designed for the measurements, comprising four one-way tetrode stages feeding a triode detector. The output indicator was a microammeter in the plate circuit of this detector, and a single low-frequency amplifier stage was included, for aural identification of incoming signals and interference. The amplifier input circuit was untuned, and the apparatus was calibrated in carrier microvolts impressed on the input circuit required to produce a suitable reading (usually $10\mu a$) upon the output instrument, for various settings of the gain control. This gain control covered a range of about 6000 to 1 in sensitivity, which was adequate for the measurements in hand. For facilitating measurements with various types of antennas, an untuned input circuit was employed.

A voltage of 20 μv applied to the input of the receiver produced a deflection of 10μa on the output instrument when the gain control was set at maximum sensitivity. A 30 per cent modulated input voltage of this magnitude produced a potential difference of 8 volts across the telephones.

The quantity measured was the high-frequency voltage impressed upon the input terminals of the receiver from an antenna of known geometry exposed to the wave field as described below. The process of taking readings consisted of manually adjusting the gain control to maintain the reading of the output instrument constant, and observing these gain control readings as a function of time or distance to the transmitting source during flights away from and toward the receiver.

THE GROUND ANTENNA

The most suitable type of antenna which can be used on a receiving set depends upon the impedance of the input circuit. With a high resistance input circuit, such as employed, the most suitable antenna has its maximum impedance when the receiver input current is a minimum: it is then very nearly half a wavelength long. Such an antenna has particularly interesting properties.

The current distribution along an open-ended straight wire is very nearly sinusoidal, as shown in Fig. 2. This figure is not quite correct, for the current never falls absolutely to zero. In the neighborhood of the anodes of current there is a very rapid change of phase of the current through 180 deg., but the amplitude of the current never falls absolutely to zero, although it reaches a very small value. This is a result of the resistance of the antenna. In the ideal theoretical case of an antenna having no resistance the current would fall to zero. The so-called effective height of an antenna is equal to the integral of the current along its length divided by the input current. Many writers prefer to refer the effective height to the maximum current in the antenna, but there appears to be very little reason for this, as any apparatus connected to the antenna is independent of the fact that the current reaches a mathematical maximum. It is only concerned with the constants of the antenna measured at the point at which the

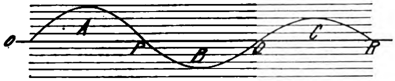

Fig. 2—Current distribution along a straight wire.

apparatus is connected. We shall therefore refer the constants of the antenna to the current flowing from it into the receiver. In this case, the effective height of the antenna is approximately equal to

$$\left[\int_0^l \sin(2\pi x/\lambda)\,dx\right]/\sin(2\pi l/\lambda)$$

It will be seen that this becomes infinite when $l=\lambda/2$. That is so because we have not taken into consideration the effect of the resistance of the antenna. This is a more complicated problem than appears at first sight, for the resistance of the antenna per unit length is by no means constant. The radiation resistance will vary considerably with the length of the antenna.

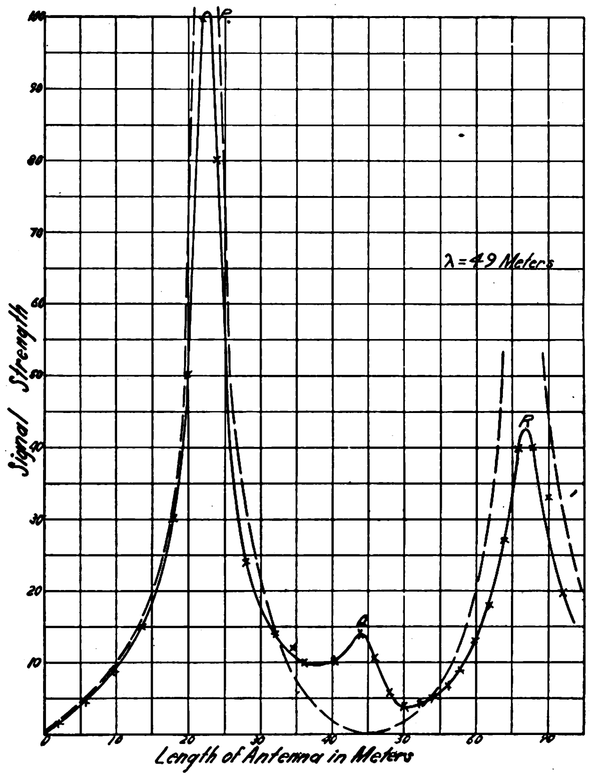

Fig. 3—Variation of effective height of antenna with length $\lambda=49$ meters.

The variation of effective height with the length of a horizontal antenna was investigated experimentally by finding the effect of the variation of the length of the antenna on the signal received from a steady source situated at a distance which was large compared to the length of the antenna. The results are shown in Fig. 3. The antenna used was horizontal and set at a height of 3 ft. above badly conducting ground. The receiver was connected at one end of the antenna and kept at a height of 3 ft. above the ground. No ground connections were used. The full line shows the experimental result, while the dotted line is the curve

$$\left[\int_0^l \sin(0.9\cdot 2\pi x/\lambda)dx\right]/\sin(0.9\cdot 2\pi l/\lambda)$$

The factor 0.9 occurs probably because of end effects and the difference in the speed of propagation of electromagnetic waves along a wire in the above conditions and *in vacuo*.

The very large effective height obtainable is noteworthy. The agreement with the theoretical calculations is very good except near the nodes of current as was to be expected. The peak P corresponds to an antenna of length OP. (Fig. 2) The area A is greater than the area B, so that when the length of the antenna is equal to OQ the signal is not zero. The fact that the effective height reaches a mathematical maximum at Q shows that the decrease in current as the point Q is approached more than counterbalances the decrease in the effective current area. The maximum at the point Q is small, however, as is to be expected. The next maximum R is large, though not so large as the first maximum. If the length were increased still more the effective height curve would be found to consist of a series of alternately large and small maxima. From the experimental curve, if we take the effective height of a small antenna to be equal to half its length we deduce that the maximum effective height is of the order of 55 meters.

A vertical antenna was briefly investigated in the neighborhood of the optimum condition by adding short horizontal lengths. This method is very approximately correct, because the e.m.f. induced in the antenna in the neighborhood of a node of current produces very little effect on the total effective e.m.f. The results were of the same order as those obtained for the horizontal antenna.

From the nature of the curve obtained it is evident that the controlling factor of the effective height at the optimum length is the current at the foot of the antenna rather than the integral of the current along the length. It can be shown by applying the simple transmission line theory that the ratio of this current to the maximum current flowing in the antenna is very approximately proportional to the resistance of the antenna (measured with respect to the maximum current). The effective height of the antenna of optimum length will therefore be inversely proportional to the resistance, so that the signal received will be proportional to

$$E/(R_a+R)$$

where $(R+R_a)$ is the total resistance of the antenna, R_a is the radiation resistance, and E is the electric intensity of the oncoming wave. When the antenna is a small fraction of a wavelength above the ground the radiation distribution from it will be very nearly of the same shape for all heights above the ground. In these conditions E^2 is proportional to R_a. If R_a is large compared to R, as will occur, when the height of the antenna above the ground is not very small, the signal will become smaller the greater the value of R_a, that is, the higher above the ground the antenna is laid, the weaker the signal. Very near the ground R begins to *preponderate*, not only because R_a is smaller but also because of increased losses; the signal in that case should decrease as the antenna is lowered. These two effects may counteract each other. It was actually found that increasing the height of the antenna from 0.65 m to 1.7 m produced only a 56 per cent increase of the signal strength when the antenna was stretched over ground of high conductivity, both for the sky ray, which must arrive at a high angle of elevation, and for the direct ray, which was generally at an angle of elevation of but a few degrees.

It must be realized that the signal received is produced by the resultant of the oncoming wave and the reflected wave on the ground. A signal is received only by virtue of the fact that these two waves are not of the same amplitude and phase, owing to the imperfect conductivity of the ground, and also because the finite distance between the antenna and its image in the ground produces a difference of phase between these two waves. On highly conducting ground, the phase difference between the two waves produced by the distance between the antenna and its image is probably the predominating factor in producing a signal in the antenna, while on badly conducting ground the most important factor is probably due to the poor reflecting properties of the ground. Over badly conducting ground one would expect, therefore, that the radiation properties of the antenna would be practically independent of the height of the antenna above the ground. This was found to be the case. The height above the ground for these conclusions to hold must never be more than a small fraction of a wavelength.

PROPAGATION MEASUREMENTS

The variation of signal strength with distance can be divided on any one run into two and sometimes three very definite stages. In the first stage, the signal received is produced by the direct ray. There is no fading, and the signal strength varies rapidly with the distance. In the last stage the sky ray is very much stronger than the direct ray, there is strong fading but the variation of the average signal strength with distance is very slow. It may sometimes happen that the direct ray falls to inaudible values before the sky ray appears. When it does appear it comes very suddenly. From a barely audible signal the strength rises to a strong signal in the course of only four or five miles. This phenomenon is a well-known characteristic of the sky ray. The intermediate stage that occurs between the point when the direct ray is too weak to be heard and the sky ray has not appeared is commonly known as the *skip distance.* It is evidently not an actual distance, for although the sky ray makes its appearance at a fairly definite distance depending on the wavelength, the line of demarcation where the ground ray stops is very indefinite. This line depends entirely upon the power of the transmitting apparatus and the sensitivity of the receiving system.

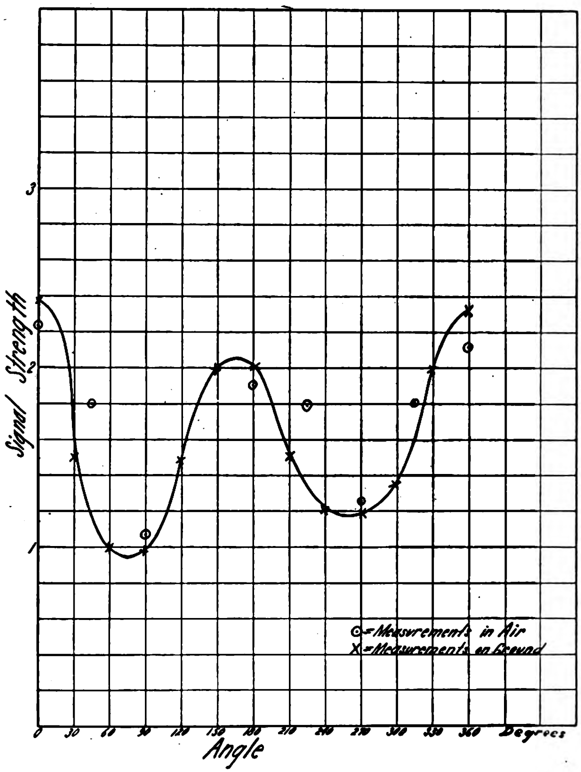

Fig. 4—Radiation distribution diagram in horizontal antenna on airplane.

The curves to be shown were not all obtained under quite identical conditions, the antennas used for reception being slightly different. For the purposes of comparison, the values obtained were adjusted for the case of an antenna of optimum length using a reduction factor deduced from the curve shown in Fig. 3. Some twenty flights were made under various conditions.

The radiation distributions of the airplane antennas in the horizontal plane were found both on the ground and in the air. Measurements were made as the airplane flew in different directions over a given point some five miles distant. In Fig. 4 is shown the result for the horizontal antenna. The crosses correspond to the points found with the airplane on the ground, while the circles correspond to the readings made in the air. It will be seen that the two sets of readings lie fairly on the same curve, a satisfactory result, considering the difficulty of flying the airplane accurately over a given point and in a given direction. The 0-deg. position corresponds to the airplane facing the ground station, positive angles being measured as the airplane is rotated in a clockwise direction looking from above it. The radiation distribution in the horizontal plane of the vertical antenna was found to be accurately circular.

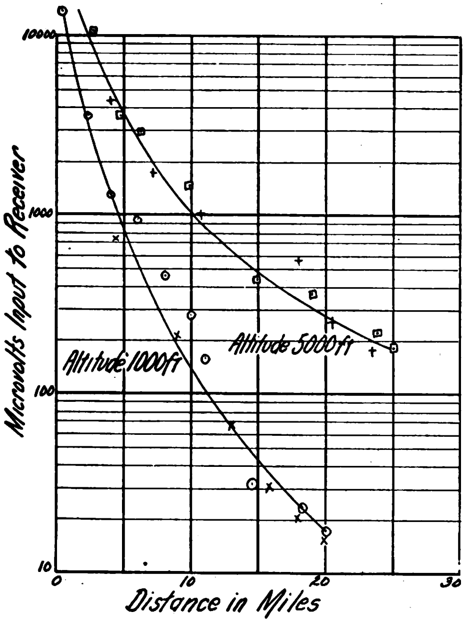

Fig. 5—Variation of strength of direct ray with distance, using horizontal antenna on badly conducting ground.

The first experiments referred to the direct ray. In the appendix a method supported by experimental evidence is given by which the effect of the altitude of the airplane can be calculated, when the variation of signal strength with distance is known at a given altitude. This refers essentially to the direct ray, for the altitude has practically no effect on the sky ray.

Typical curves for the direct ray are shown in Fig. 5. This figure shows two curves corresponding to altitudes of 1000 ft. and 5000 ft., respectively, using a horizontal antenna over poorly conducting ground. The crosses correspond to readings taken with the antenna 2 ft. above the ground, while the enclosed dots are the readings taken with the antenna 15 ft. above the ground. It will be seen, as mentioned above, that the height of the antenna above the ground does not materially affect the signal strength.

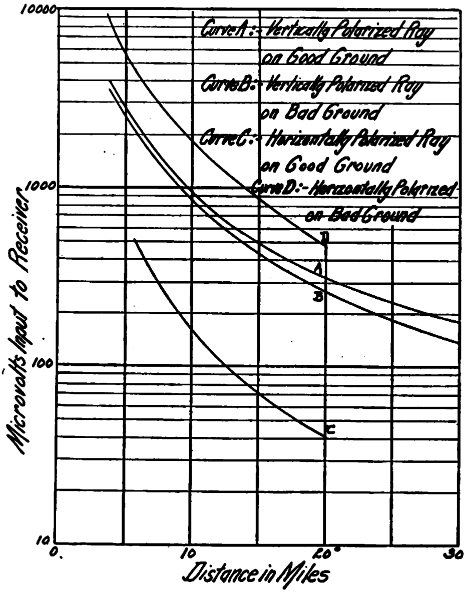

Fig. 6—Variation of signal strength of direct ray with distance.

Similar curves were obtained for the horizontally and vertically polarized rays over both good and poorly conducting ground. The average results of a number of flights reduced to a common altitude of flight of 5000 ft. are shown in Fig. 6. In this figure four curves are shown giving the relative signal strength of the direct ray for vertically and horizontally polarized waves on both high and low conductivity ground.

A typical curve for a longer distance flight is shown in Fig. 7. Many such curves were obtained and were chiefly remarkable for the similarity. In this figure, which corresponds to reception of the horizontally polarized ray over bad ground while flying at an altitude of 5000 ft. the two stages referred to alone are well brought out. In the first stage up to a distance of 20 miles, the signal is very steady but decreases very rapidly with the distance. In the second stage, the signal is very unsteady, varying over a large range of amplitudes in the course of a few seconds. In order to show this fading effect on the diagram, each reading covered a period of 5 to 10 sec., and the maximum and minimum values of the signal during that period were noted. These values are plotted and joined by a vertical line. It will be noted that the mean signal strength is remarkably constant over the distance investigated. Flights as long as 600 miles were undertaken; measurements were not carried out over the whole distance, but signals were received continuously on telephone and there did not appear to be any very large variation of the mean signal strength. In fact, the signal seemed to tend to increase in intensity towards the end of the flight, that is, at distances of 400 to 600 miles.

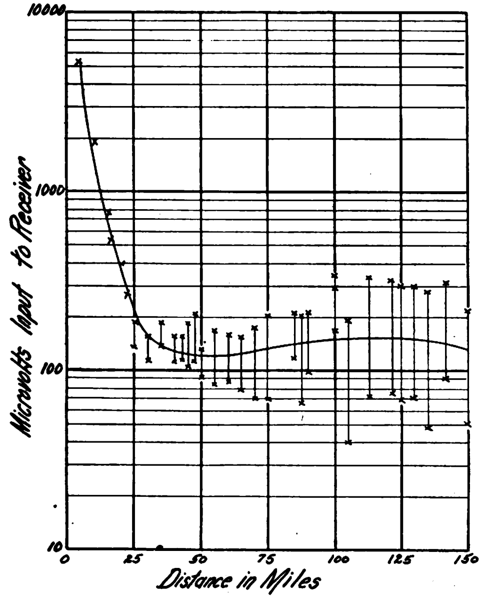

Fig. 7—Variation of signal strength with distance, using horizontal antenna on badly conducting ground.

The altitude of the airplane was not found to have any appreciable effect on the intensity of the signal received by the sky ray.

The test was repeated over high conductivity ground, and the signal was found to be considerably weaker, the mean being of the order of 25-μv input to the receiver. With the vertical antenna, the sky ray became appreciable only at a distance of 50 miles. From that distance up to 150 miles the mean signal was fairly constant in value, but its intensity was only about 20 μv. The effect of the conductivity of the ground on the signal received on the vertical antenna is small. When flying at very low altitudes, of the order of 500 ft., a skip distance is liable to occur on the vertical antenna, because of the fact that the sky ray is comparatively weak. This distance is, however, small, lying between 40 and 50 miles, and would disappear if greater power were used in the transmitter.

The explanation of these results is to be found in the difference in the radiation distribution in the vertical plane of the horizontal and vertical antennas both on the airplane and on the ground.

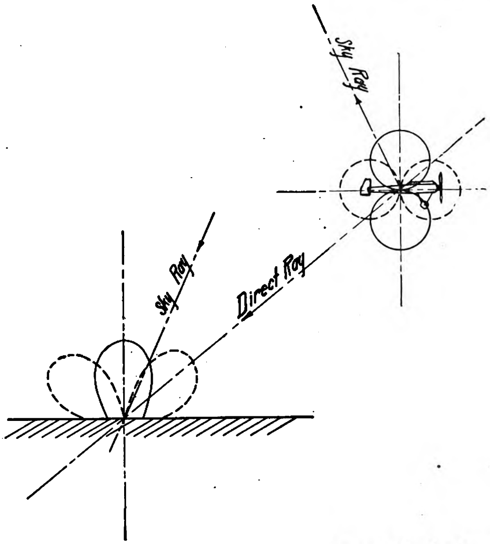

Fig. 8—Approximate radiation distribution in the vertical plane of antennas on ground and on airplane.

The direct ray is incident at the ground antenna at a very low angle of elevation. In Fig. 8 is shown roughly the radiation distribution diagram in a vertical plane of the vertical antenna (shown by the dotted lines) and of the horizontal antennas (shown by the full lines) both on the airplane and on the ground. It is readily seen that on the vertical antenna the effect of the sky ray is weak compared to that of the direct ray, while on the horizontally polarized wave the opposite is the case. This is in complete agreement with the experimental results obtained.

The effect of the conductivity of the ground is also in agreement with the theory. Over good ground the image of the horizontal antenna very nearly neutralizes the effect of the antenna itself, so that any decrease in the conductivity is likely to have a very appreciable improvement in the strength of the signal received. With the vertical antenna, the image in the ground actually helps the antenna (unless the ray is coming at very nearly grazing incidence) so that, though an improvement is to be expected when the antenna is set over highly conducting ground, the extent of this improvement is not likely to be very appreciable.

CONCLUSION

The results obtained can be conveniently summarized by means of the following table:

Ray	Type of Ground	Signal Strength (μv on input) Horizontally Polarized	Signal Strength (μv on input) Vertically Polarized
Direct Ray at 20 miles and 5000 feet altitude	Bad Conductor	500	280
	Good Conductor	40	320
Sky Ray	Bad Conductor	120	20
	Good Conductor	25	20
Sky ray becomes appreciable at a distance of		20 miles	50 miles

In reading this table it must be remembered that the figures refer to reception with a particularly effective antenna, the effective height of which is about 55 meters, and that transmissions from the airplane were carried out on an antenna only 6 ft. high for the vertically polarized and with a horizontal antenna, which, from its size, may be expected to be fairly efficient. There is little doubt that, if it were practicable to employ on the airplane a vertical antenna as efficient as the horizontal antenna, transmissions with verticaly polarized waves would be far more favorable than those with horizontally polarized waves. As it is, the preference between the horizontally and vertically polarized waves for the direct ray depends on the ground constants, while for the sky ray the horizontally polarized ray seems to be preferable. All these results refer to a frequency of 6155 kc. If this frequency were decreased, the short vertical antenna would be slightly less effective, while the horizontal antenna would lose much of its efficiency. This applies not only to the airplane antenna, but also to the horizontal ground antenna. At lower frequencies we might well expect the vertical antenna to be more suitable than the horizontal. At higher frequencies, the airplane antennas could both be improved, the horizontal only to a small extent and the vertical one quite appreciably, while the horizontal ground antenna would also be improved. Another factor would, however, enter into consideration. The sky ray would make its appearance only at a considerably greater distance, and the phenomenon of the skip distance might occur. If it is desired to work on the direct ray only to eliminate fading, specially designed antenna which will radiate very poorly at high angles of elevation will be necessary. It is gratifying to find that such intense signals can be received with suitably designed apparatus using only two to three watts in the transmitter for distances as large as 600 miles without the signal giving signs of failing, and it is even more gratifying to find that, at any rate during daytime when all the above measurements were made, no period was found during which a telegraph signal disappeared or became too weak to be read.

The results given in this paper, though obtained with very small powers, will apply in proportion to higher powers, for the sky ray will be increased in the same ratio as the direct ray. Fading will begin to appear when the direct ray falls to a value comparable with the sky ray, and, since the ratio of the two is independent of the power of the transmitter, fading will start at the same distance away regardless of the power transmitted.

In conclusion, the authors wish to thank L. M. Hull for very helpful suggestions made during the whole course of this investigation.

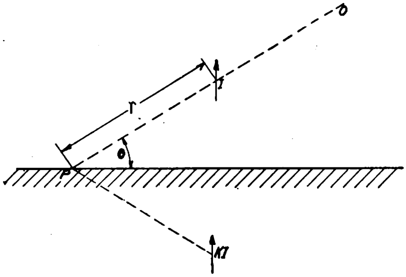

Fig. 9—Effect of conductivity of ground on image.

Appendix

REDUCTION OF AIRPLANE TRANSMISSION CHARACTERISTICS FROM ONE ALTITUDE TO ANOTHER

The method described here for reducing a signal strength vs. distance curve for a flight at a known altitude to another at another altitude is based on the theory put forward by T. L. Eckersley[1], and amplified in the discussion by one of the present authors. In this theory which has been used by S. Ballantine for computing a special case[2] it is pointed out that the image of an antenna element I is KI, as shown in Fig. 9, that the value of K is usually a complex number and depends on the angle θ and the electrical constants of the ground in the neighborhood of the point P. If this theory is correct and only the direct ray reaches the point P, the signal strength as the antenna element is moved along any radial line such as PO will be inversely proportional to the distance r of the element I. That is, when θ is kept constant, the signal strength will be inversely proportional to the distance. If the element is moved horizontally the signal will decrease more rapidly, the law tending to an inverse square law.

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[1] "Short-wave wireless telegraph," *Jour.* I. E. E. (London), **65**, 600–638; 1927.

[2] PROC. I. R. E., **16**, 513–518; April, 1928.

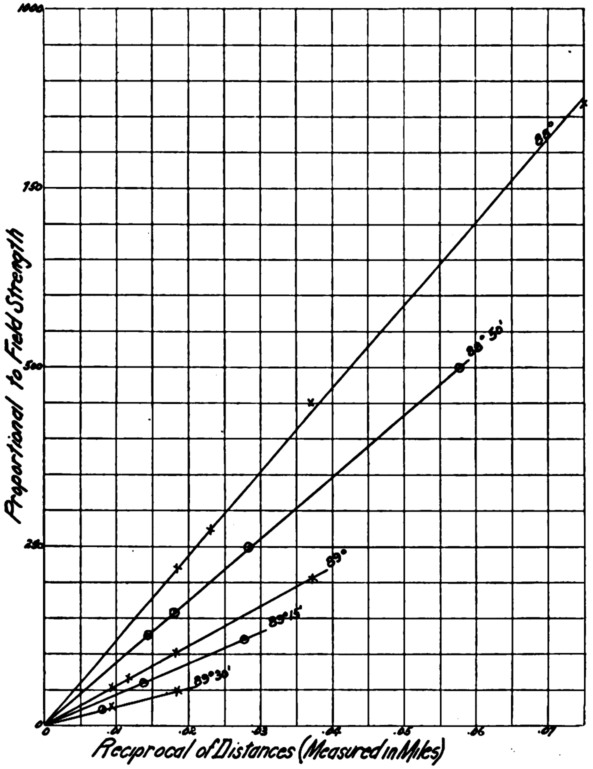

Fig. 10—Signal strength plotted against reciprocal of distance for various inclinations to the vertical.

These results were confirmed from the careful data accumulated by D. K. Martin[3]. In these experiments the signal strength was measured with the airplane traveling horizontally at various heights above the ground for distances as great as 120 miles from the ground station. From the data published Fig. 10 has been obtained. This shows the variation of field strength, with distance for various inclinations to the vertical of the line joining the airplane to the ground station. It will be seen how accurately the values lie on straight lines even for such smaller angles of elevation as 1/2 deg. as the theory outlined would lead one to expect.

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[3] "Laying the foundation for aircraft communication," *Bell Laboratories Record*, 7, 315–318, 1929.

An examination of Fig. 5 leads to an identical result. From this inverse distance law, it is not difficult to reduce any curve for a flight at a given altitude to one at another. Suppose we have the curve for a flight at an altitude of 1000 ft. and we require to know the result at an altitude of 5000 ft. The signal strength at a distance of five miles at 1000 ft. will be five times the signal strength at a distance of 25 miles at 5000 ft., and so on for other distances.

The fact that this law appears to hold with such accuracy would tend to show that it is only the constants of the ground in the immediate neighborhood of the ground station that produce attenuation of the waves, in excess of the inverse distance law, for otherwise one would expect the constants of the ground at a distance from the ground station to become effective as the distance of the airplane from the ground stations is increased, even though the angle θ remains constant.

This inverse distance law holds, of course, only for the direct ray, and in reducing curves from one altitude to another the reduction must be applied only to that part of the curve for which there is no fading. When fading occurs, that is when the sky ray becomes predominant. The altitude of the airplane produces no effect on the signal strength.

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Proceedings of the Institute of Radio Engineers
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TEMPERATURE RATING OF WIND-DRIVEN AIRCRAFT RADIO GENERATORS*

By
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Summary—*Measurements of temperature rise of generator under load are given for still air and in flight. Theory of heat emission is briefly reviewed, constants of emissivity are given from which rising and final temperatures of generators can be approximated. Conditions governing the rating of wind-driven aircraft radio generators are discussed.*

In the past the usual source of power for aircraft radio transmitters has been the wind-driven generator, which is mounted on the wings or other structure of the plane, usually in such a position that it comes in the slip stream of the main propellers. Exact knowledge of the cooling conditions which exist in generators operated in this manner has not been hitherto available, which fact has led to some uncertainty in their ratings.

In order to throw light on this condition it was undertaken to determine the relative temperature rise in such generators in flight and in still air. As a means of measurement thermocouples were applied to various points of the generator under test. Through the kindness of Dr. E. G. Lunn, of the Naval Research Laboratory, a number of group thermocouples were made up for use with a standard Leeds and Northrup temperature bridge. Each group consisted of six elements in series, each element being made of No. 28 iron and constantan wire. The cold junctions were in glass tubes to permit use of ice in thermos bottles to fix temperature at cold junction. The apparatus for use in flight is shown in Figs. 1, 2, and 3 as installed in a standard O2-U land plane.

A generator selected for test was given a run in the laboratory at three values of loading during which temperature was recorded as shown in Figs. 4, 5, and 6. The generator was then installed on plane and temperature runs made in flight with the same load conditions, as indicated on Figs. 7, 8, and 9.

To review briefly the physical conditions which govern temperature rise and heat emission, heat develops in the structure of a generator at various points in proportion to the losses. These may be I^2R losses in the windings of the armature and field, or a result of eddy currents,

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* Dewey decimal classification: 621.313.23. Original manuscript received by the Institute, October 7, 1929.

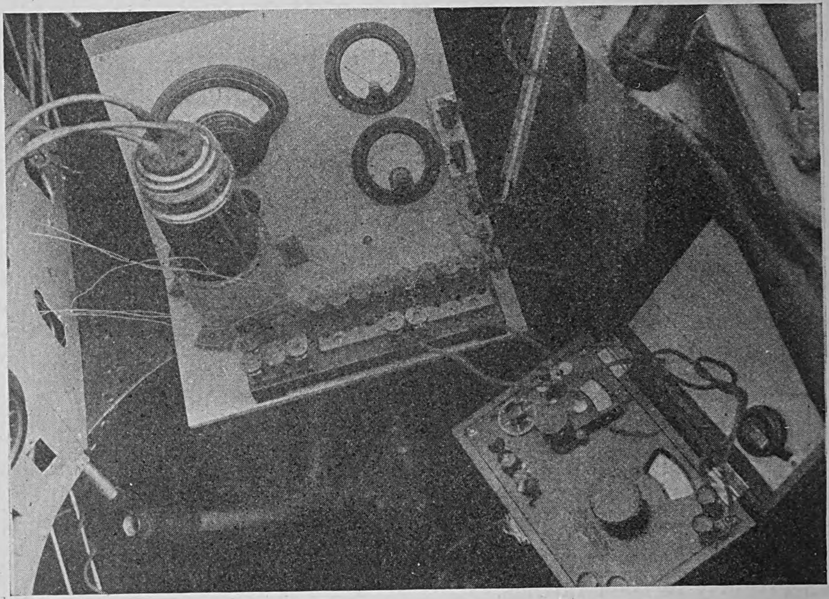

Fig. 1—View down into cockpit showing temperature bridge, instruments, and switching arrangement.

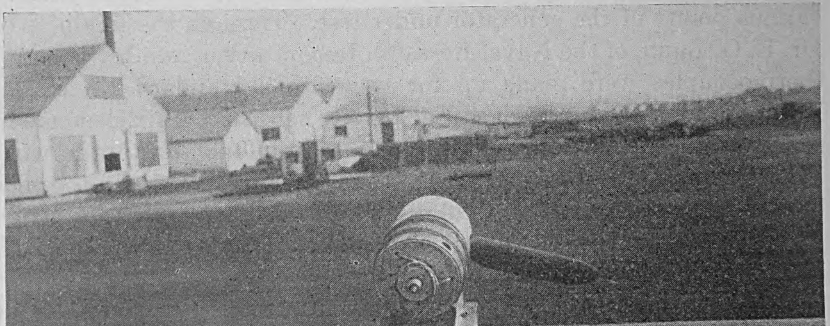

Fig. 2—Generator mounted on upper wing showing leads to thermocouples.

hysteresis, losses in the iron core, friction losses, etc. From the point of formation, the heat flows toward the surface of emission following the laws of conduction and is there emitted. The temperature gradient between the points of heat generation and the emitting surfaces is proportional to the length of the path, the conductivity of the material, and is inversely proportional to the cross section of the path.

At the surface the heat is emitted in three ways. A certain amount passes by conduction to the supports of the machine. A certain other amount passes directly to surrounding bodies by radiation. This is proportional to the difference in fourth powers of the absolute temperatures of the heat emitting and the heat receiving bodies. It is independent of the temperature of intervening air, but varies greatly with the nature of the heat emitting surface. A certain other amount of heat is taken up by convection in warming the air which comes in contact with heat emitting body. Dr. Langmuir[1] has shown that in still air the emission by convection about equals the heat flow, which would be occasioned through an air film 4.3 mm in thickness by the temperature difference between hot body and the cooling air. The

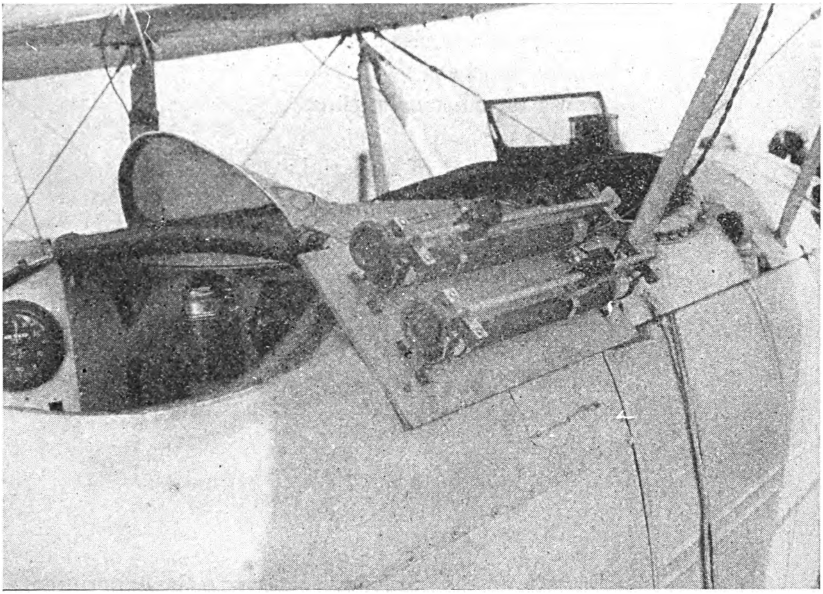

Fig. 3.—Rheostats for generator load.

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[1] Laws of heat transmission in electrical machinery, *Trans.* A.I.E.E., **XXXII**, 301.

emission by convection is probably independent of the nature of the emitting surface.

The heat given out in these three ways comprises the total emission, and is related to the dimensions of the cooling body by Newton's well-known law of emissivity, which may be stated as follows:

$$T = T_f - T_r e^{-(Ca/mh)t}. \tag{1}$$

Where:

T = temperature at any time
T_f = final temperature
T_r = temperature rise
e = naperian base
a = area of emitting surface
m = mass
h = specific heat of material
t = time
C = constant of emissivity

When final temperature is reached a state of equilibrium exists, in which the heat emitted equals the heat supplied from losses.

$$CaT_r = W$$

or

$$T_r = \frac{W}{Ca} \tag{2}$$

An attempt has been made to apply this theory to the experimental curves here given with a view to establishing constants for the predetermination of temperature rise. In making the preliminary attempt several assumptions are made.

(a) Owing to the small mass the temperature rise in test generator will be relatively fast.

(b) Owing to the small dimensions, the temperature gradient between internal points and surface, as equilibrium is approached, will be relatively small.

(c) Under flight conditions the very intense convection will take up the greater part of the emission. The heat transferred by conduction will be small as the mass and conductivity of generator supports are small.

Considering the generator under test as a cylinder omitting the stream line tail piece, dimensions were found as follows: weight in pounds, 18.75; weight in grams, 8500.; area in square inches, 161.9; area in square cm, 1042. Using these values in (1) and using the temperature limits as observed in Fig. 4 a value of C was determined as 0.0287. Using this value computed values were found as shown by

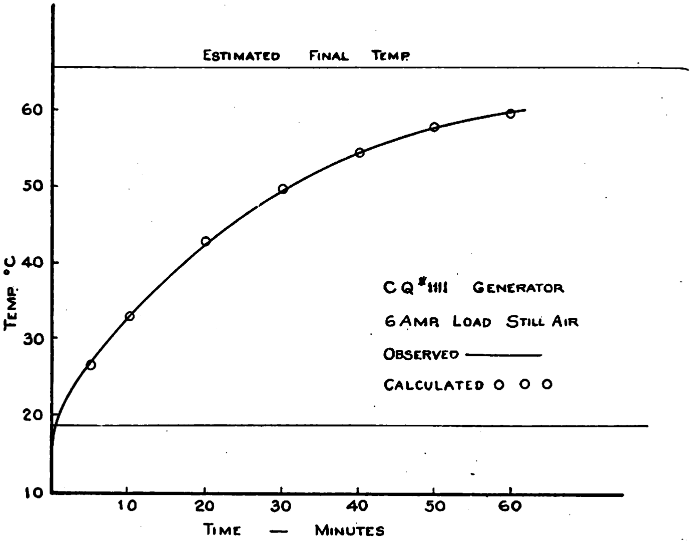

Fig. 4

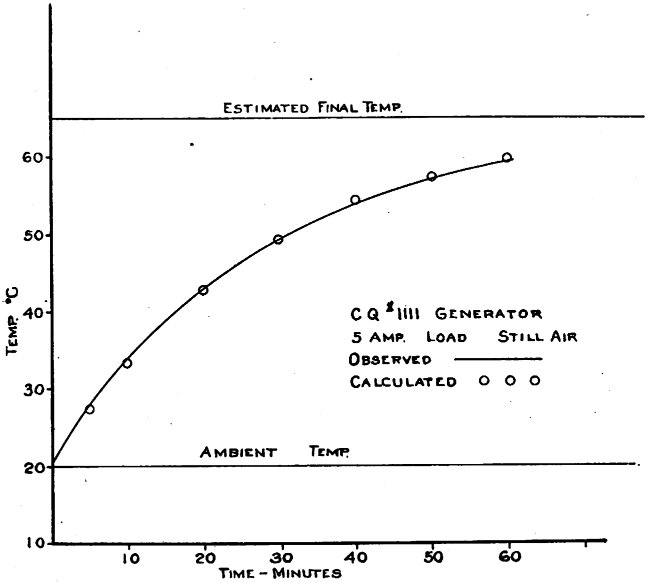

Fig. 5

circles in Figs. 4, 5, and 6. The close agreement with the observed curves is obvious.

Equation (2) indicates means of estimating the final temperature which generator will reach at which heat emitted equals the heat generated by losses. By means of a transmission dynamometer, the overall efficiency and losses in the machine were determined.

	Losses in Watts	6-ampere load	5-ampere load	4-ampere load
A-C armature		18.	12.5	8.
Friction and windage	28.8			
Brush friction	18.			
Field loss				
Copper	26.			
Iron	44.			
Exciter arm	3.2	120	120	120
Total watts lost		138	132.5	128
Total gram cal. per minute		1980	1915.	1835

One watt is equivalent to 14.33 gram calories per minute. The specific heat of generator is taken as 0.1. Temperatures are in degrees C.

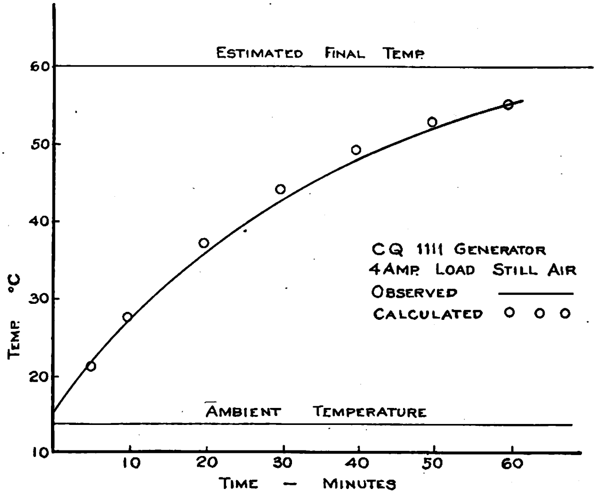

Fig. 6

Weight is used for mass, and constants found should be interpreted accordingly.

Final Temp. rise	6-ampere load	5-ampere load	4-ampere load
For still air computed from equation (2)	66 deg.	64 deg.	64 deg.
Observed, Figs. 4, 5, and 6	46 "	45 "	46 "

The calculated final temperatures resulting by this procedure were higher than the observed curves of Figs. 4, 5, and 6 would indicate. However, a value of 0.04 for *C* when used in (2) with losses as above measured gives a final temperature which agrees closely with the observed data.

Extending this method to observations on several other machines, both in still air and in flight, average values for emissivity constant were found as follows: in still air, 0.033; in flight, 0.0925. It is found that both rising and final temperatures agree with the computed temperatures derived from these constants within about 20 per cent, sometimes much closer.

Several unusual features affect the rating of wind-driven aircraft radio generators. The load during transmission of radio code is intermittent as the key is open about 1/3 of the time even during sending. The periods of use in aircraft service are short, a 10-minute message being a rather long one. The temperature of surrounding air normally decreases with the altitude of the plane at about 2 deg. per 1000 ft. All these conditions are favorable. Against this is the effect of the reduction in atmospheric pressure on cooling conditions. At first this might seem important since the cooling effect of rarified air is materially reduced, a fact which is taken account of by builders of stationary machines which are to be used at high altitudes.[2] It must be remembered, however, that wind-driven aircraft generators are usually mounted in the direct blast from the slip stream of the main propellers. As the air becomes rarified as a result of increasing altitude, the driving propeller of the plane turns faster, forcing a greater volume of air past the wings. While the volume is greater the mass tends to remain constant, for it is the action of the *mass* of air passing the wings which keeps the plane up. The cooling effect on the generator

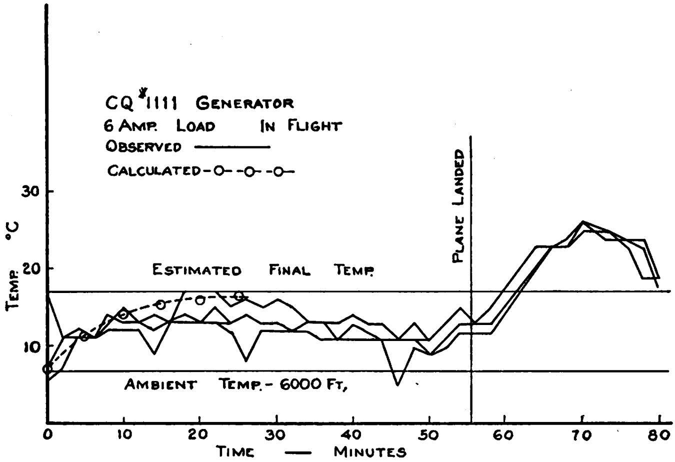

Fig. 7

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[2] Carl Fechheimer, *Trans.* A.I.E.E., p. 124, February, 1926.

is also proportional to the mass of air, and as long as sufficient air is passed to keep the plane in flight the cooling cannot be seriously decreased. This is a different condition from that which prevails when a stationary generator at high altitude is cooled by exhaust fans.

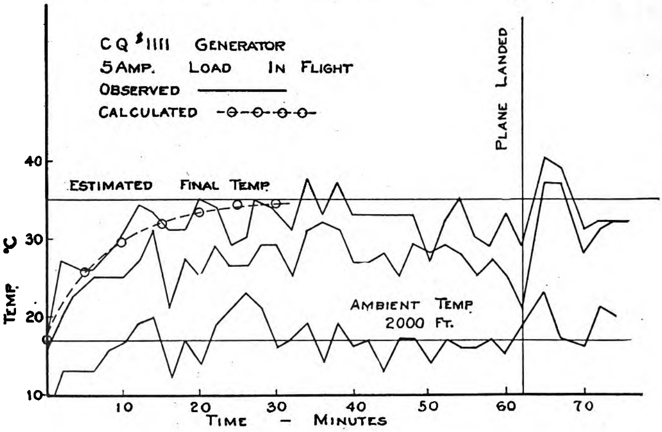

Fig. 8

Then the volume of air driven through ducts tends to remain constant, while the mass diminishes.

As a conclusion, from the preceding work it seems evident that the cooling of aircraft wind-driven generators in flight is about three times as great as it is in still air. Because of this and the favorable conditions of operation just cited, it is believed that in the past such generators have been consistently underrated. One point, however must be taken into consideration. It is evident from Fig. 7 showing temperature measurements in flight that there was a sudden rise

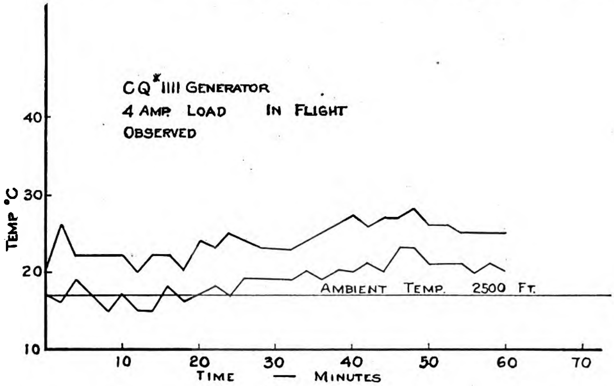

Fig. 9

of temperature immediately after the plane landed. This accomplished the sudden stoppage of forced cooling, which permitted a momentary increase in temperature of 30 per cent over that prevailing during flight. A similar momentary rise in temperature also takes place in the adjacent motor of the plane, which may have some reaction on the generator. Though of short duration, the effect must be taken into account.

It is desired to acknowledge the valuable cooperation of Assistant Radio Inspector A. H. Johnson, of the Naval Research Laboratory, who made the installations and took the measurements in flight.

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APPLYING THE RADIO RANGE TO THE AIRWAYS*

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Summary—*In pursuance of the program of radio aids to flight undertaken by the Department of Commerce, experimental and installation work has progressed in the application of the radio range (directive radiobeacon) along the airways. To date, only the aural type has been put into routine daily operation, and it is the only type of beacon considered in this paper, which discusses methods of adjusting the space pattern of the beacon system in order that the courses may align with the fixed airways. These beacons also need to be readily distinguished from one another, and so designed that a minimum of interference is met.*

By using a vertical wire antenna in addition to the loop antennas and varying the relative power in the two loop antennas it was found possible to secure practically any array of courses desired. The radio ranges at Hadley Field, N. J., and Bellefonte, Pa., were employed in the experimental work with excellent results. The field-intensity measurements made gave space patterns which checked very well with the theoretical patterns for such antenna systems.

Careful spacing of the radio ranges within the frequency band, as well as distinctive coding of each beacon, solved the problem of interference.

In selecting the proper coding for the beacons a study of the physiological effects of various sound groups was made, and the final coding chosen was such as to give a signal of equal time duration on each side of the course. The signals still interlock to give the customary long dash when on the course.

The technique of applying these modifications was developed to the point where the adjustments could be made by the installation crew in the field without the aid of an engineer.

INTRODUCTION

THE aural type of radio range[1] was installed by the Department of Commerce at Hadley Field, N. J., Bellefonte, Pa., and Cleveland, Ohio, in 1927 and 1928. After the first tests, it became necessary to adjust the beacon system in order to make the courses provided coincide with the fixed airways. The original aural type radio range established an equisignal zone of about 3 deg. width extending in a straight line through the beacon approximately 100 miles either way. Another equisignal zone, at right angles thereto, approximately 12 deg. in width, was also present, but on account of

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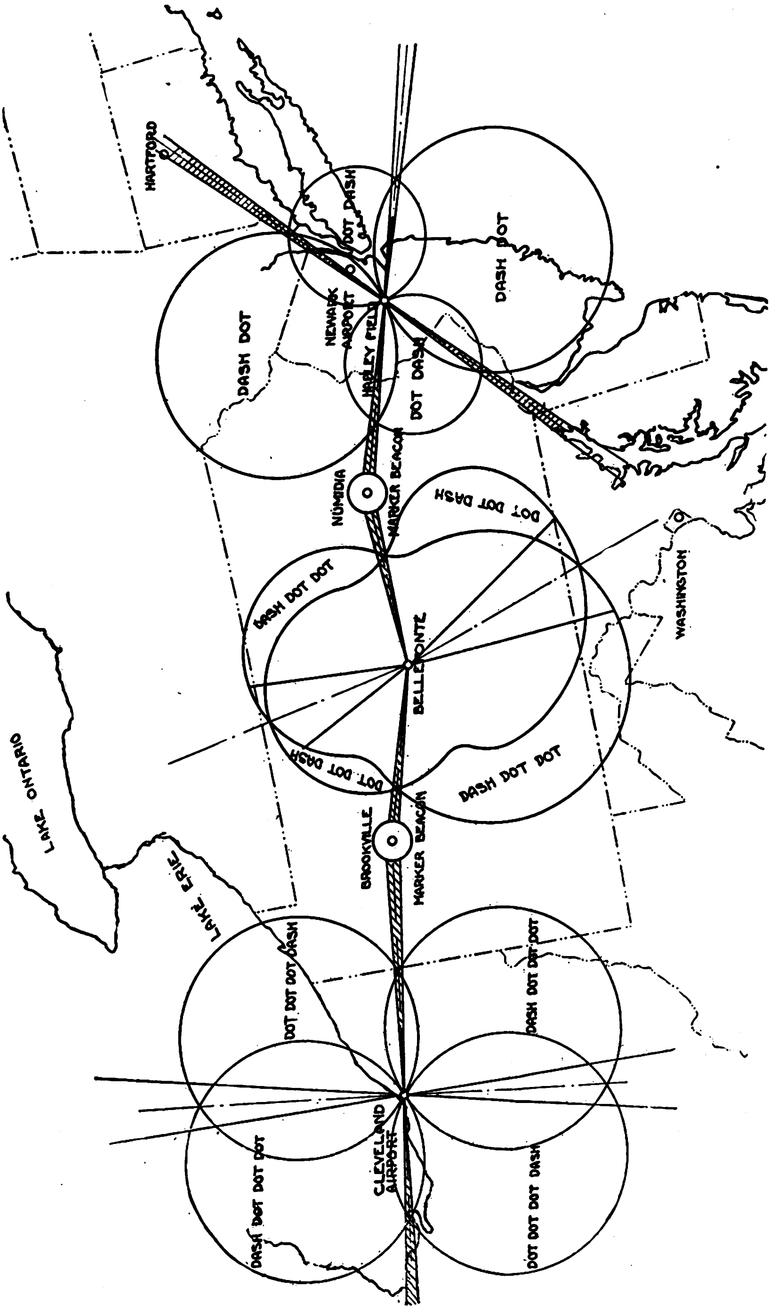
*Dewey decimal classification: R526.1. Original manuscript received by the Institute, October 7, 1929. Publication approved by the Director of the Bureau of Standards of the U. S. Department of Commerce. To be published in a forthcoming issue of the Bureau of Standards Journal of Research.

[1] F. H. Engel and F. W. Dunmore, "A directive type of radio beacon and its application to navigation," Scientific Paper, Bureau of Standards No. 400, January, 1924. Murphy and Wolfe, "Stationary and rotating equisignal beacons," *Jour. S.A.E.*, **19**, 209; September, 1926. J. H. Dellinger and Haraden Pratt, "Development of radio aids to air navigation," Proc. I. R. E., **16**, 890; July, 1928.


HARTFORD
DOT DASH
DASH DOT
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NEWARK AIRPORT
HADLEY FIELD
DOT DASH
NUMIDIA
MARKER BEACON
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BELLEFONTE
WASHINGTON
DOT DOT DASH
DASH DOT DOT
BROOKVILLE
MARKER BEACON
LAKE ONTARIO
LAKE ERIE
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CLEVELAND AIRPORT
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Fig. 1—Air route, Cleveland to New York.

its excessive width was not usable. An airway extending 200 miles in a straight line with a field at its center is a situation seldom encountered. The use of lower power beacons at more frequent intervals, and between fields, might solve this difficulty, but would raise problems of interference and economy. In the actual use of the beacons along the airways the following problems had to be solved.

(a) Interference had to be kept to a minimum in order to prevent confusion.

(b) The course outlined had to be easily recognizable so that a slight deviation might be noticed.

(c) Radio ranges had to be so designed that they could be located at bends in the airway route; that is, the course given by the beacon must not be a straight line through the beacon but an obtuse angle of controllable magnitude. Furthermore, both legs of the angle must be capable of simultaneous operation in order to accommodate increasing air traffic.

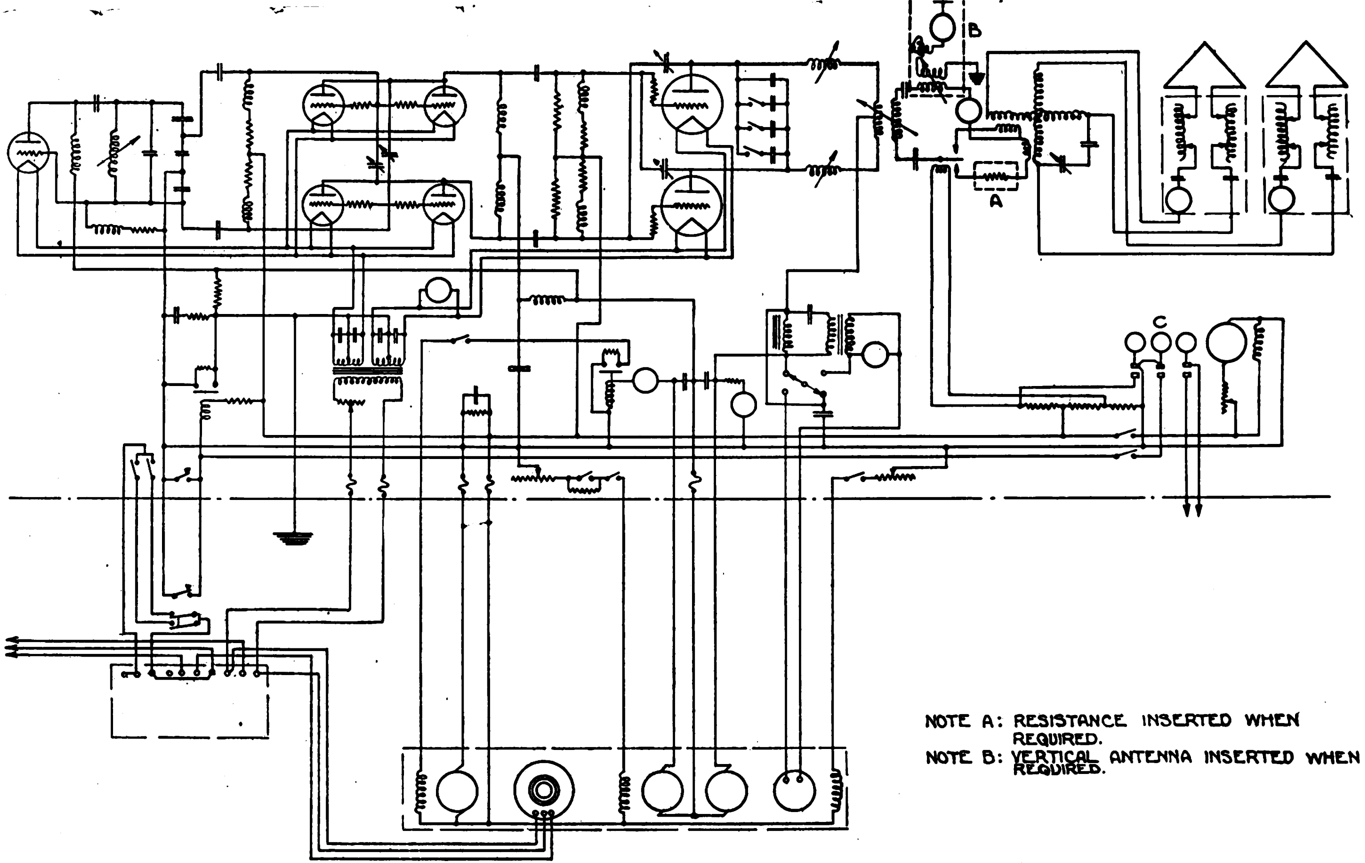
(d) Finally, it was necessary to service more than one course from a single radio range, a further development of the previous idea. Terminals in larger cities have many air routes entering, and each of these routes should be outlined by a beacon course. Since it would not be economically possible to install a separate beacon for each route the previous system had to be modified, to accommodate multiple routes.

These conditions have all been satisfied on the radio range system as now in use on the airways, if the number of courses to be served does not exceed four. The purpose of this paper is to give a detailed account of the means employed to meet these conditions.

I. INTERFERENCE BETWEEN RADIO RANGES

Two methods have been successfully adopted to differentiate between radio ranges. The first method was to allocate different frequencies in the band from 285 to 315 kc to each radio range. Particular care was taken to separate adjacent beacon frequencies as much as could consistently be done in the limited available frequency range. In no case were adjacent radio range station frequencies selected closer than 8 kc. When utilizing the ordinary aircraft receiver using two stages of shield-grid tube, radio-frequency amplification, detector and two stages audio with a 6-ft. rod antenna, it was possible to tune in either a beacon 100 miles ahead or 100 miles behind, when both were operating simultaneously on frequencies separated by 8 kc. The average power output from a beacon installation is 2 kw.

The second method used to eliminate possible confusion between radio ranges was to establish a system of distinctive code signals.



NOTE A: RESISTANCE INSERTED WHEN REQUIRED.

NOTE B: VERTICAL ANTENNA INSERTED WHEN REQUIRED.

Fig. 2—Schematic diagram of radio range. The master oscillator is a 50-watt tube employing the Colpitts circuit. This is amplified by a set of four 50-watt tubes operating in a push-pull circuit and cross-neutralized. The final power amplifier contains two 1-kw tubes of which the plate circuit is tuned. The tuned plate circuit is coupled by an untuned link circuit to the goniometer primaries and keying is accomplished at this point. The keying cams are shown at *C*. One cam controls the input power and another connects the primaries of the goniometer alternately to the circuit. The resistance at *A* permits shifting the 90-deg. courses as desired and the vertical wire antenna shown at *B* introduces the shift of the 180-deg. courses. Either or both may be employed as desired.

The signals of necessity had to be such that they interlocked and gave the resulting long dash when on the course. The first group of signals adopted were (—. and .—); (—.— and ..); (—..— and ...); and (— — and ..); two characteristic signals being required for each interlocking radio range. Results showed that radio ranges using the —. and .— characteristics gave very distinctive courses. However, when the characters —.— and .. were used as interlocking signals, a peculiar physiological offset was observed. Tests indicated that the course was apparently bent slightly in an *S* shape. In reality, field-strength measurements proved the course to be straight. When close to the range and on the course as indicated by field-intensity measurement, the character —.— would apparently predominate, causing the pilot to bear toward the .. side of the course. When at approximately 100 miles from the range, tests indicated that the .. character would predominate while on the course, thus causing the pilot to compensate for this effect by flying on the —.— side of the course. As a result the pilot does not fly on the true course but flies on a curved course which does not coincide with the airway.

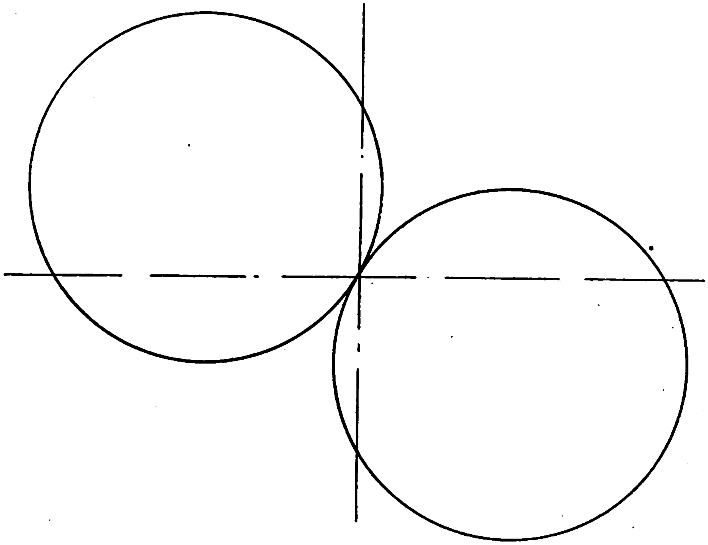


Fig. 3—Theoretical field-intensity pattern of a single phantom loop antenna without the use of a vertical antenna.

In order to eliminate this confusing effect it was found necessary to use characteristics that had equal time durations and were the inverse of each other. This plan was adopted since the —. and .— were found to give such excellent results. The characteristics chosen were

(—.. and ..—); (—... and ...—). These have been used for some time and have eliminated difficulties encountered with dissymmetrical arrangements, but are not favored by the pilot.

Recently the Airways Division changed signals to —. and .— combinations, transmitted in groups of 2 to 12 signals per cycle. By counting the number of signals the station may be identified, this system being similar to a clock tolling the hour. In addition, beacons are identified by station announcement every 15 minutes.

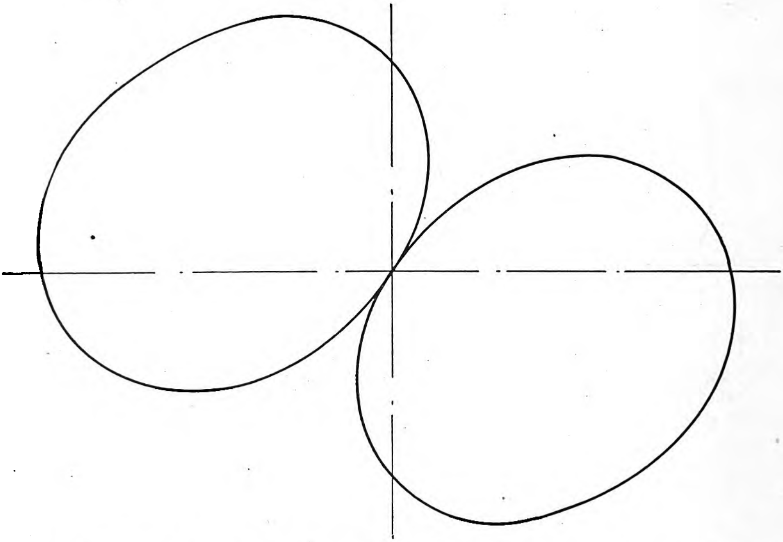


Fig. 3a—Observed field-intensity pattern of a single phantom loop antenna without the use of a vertical antenna.

II. Alignment of Courses with Airways

It has long been appreciated that a limitation of the radio range in its early form has been the fact that the four courses emanating from a single radio range have been at fixed angles of 90 deg. to each other. It is obvious that such an arrangement could serve only one course at a time situated at an airport except where by rare coincidence two courses extended at 90 deg. or 180 deg. fron an airport. In order to make at least two of the four available courses useful, the radiobeacon installations could be situated half-way between points on the straight airways at distances of approximately 200 miles. This would make it possible to use two of the four courses, and at the same time to reduce the angular width of the equisignal zone since the two courses were displaced 180 deg. This is accomplished by increasing the angle between the two primary coils of the goniometer to about 120 deg. instead of 90 deg. See Fig. 4. Some installations of this type were made along the Transcontinental Airway.

Placing radio ranges in this manner has one particularly annoying feature in that the course is usually broadest where it should be narrowest. To make this statement clear, it should first be understood that an equisignal course is a true angular function, i.e., directly over

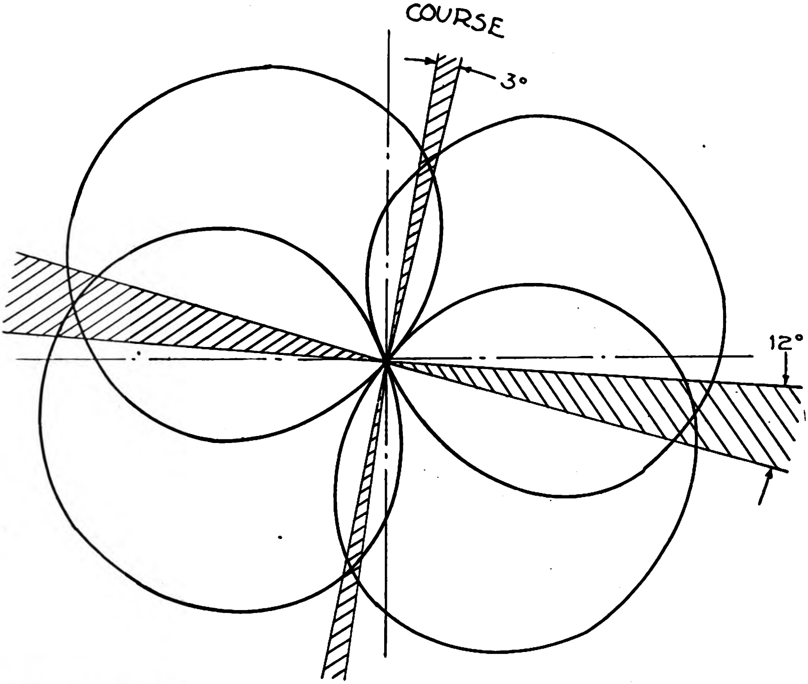


Fig. 4—Observed field-intensity patterns of each phantom loop antenna energized separately; goniometer primaries set at 120 deg.

the center of the loop antenna the courses have an infinitesimal width, and gradually the width increases to approximately 6 miles at a hundred mile distance from the radio range. When a pilot flies from a point at a distance from the radio range, it is evident that his course gradually becomes sharp and narrower as he approaches the beacon installation, and if he keeps on his course it will take him directly over the radio tower. This feature alone is one of paramount importance and cannot be neglected, since experience has shown that when the radio range is located adjacent to the terminal airport, pilots are able to locate the airport and make safe landings when conditions are such that they would otherwise be unable to find the terminal airport. Hence, when a radiobeacon is located at a terminal airport there are two very distinct advantages: first, the course becomes narrower as the pilot approaches the beacon; secondly, the beacon informs the pilot definitely when he has passed over the tower. Therefore, it practically fixes a point in space directly over the antennas. It is apparent that such a point in space would be of unquestionable value if located adjacent to a terminal airport instead of halfway between two airports. It was with these particular features in mind that the problem of fitting radio ranges to the airways was attacked.

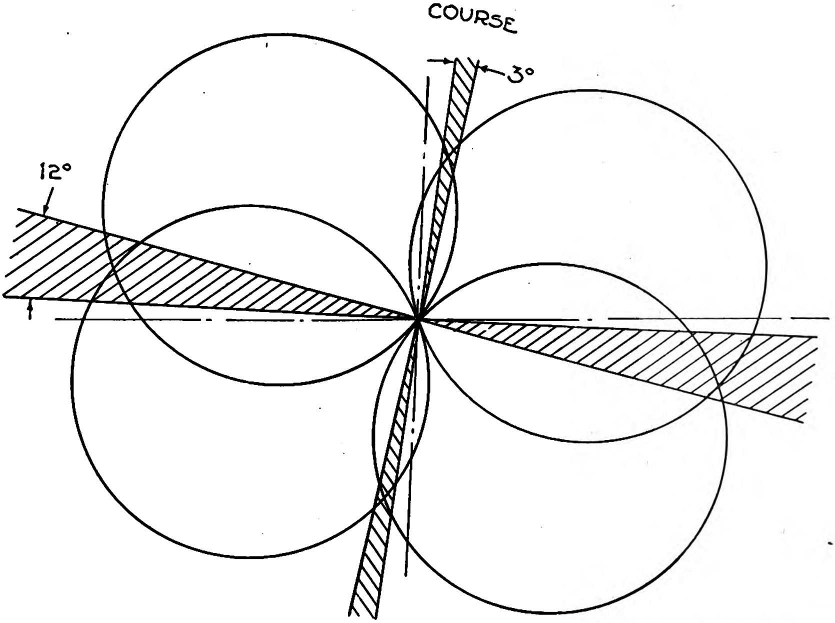


Fig. 4a—Theoretical field-intensity patterns of each phantom loop antenna energized separately; goniometer primaries set at 120 deg.

Two fundamental schemes have been developed which permit the use of two fixed courses displaced at any angle varying between 45 deg. and 180 deg. Courses emanating at angles less than 45 deg. cannot practically be serviced with one installation, because an excessive decrease in power results from such an attempt.

Method Used for Courses Emanating at Angles between 45 Deg. and 135 Deg. The method used to serve courses emanating from an airport at angles varying between 45 deg. and 135 deg. is as follows: a non-inductive resistance is inserted in the link circuit in series with oneof the goniometer primaries. This reduces the power in one phantom loop antenna[2] while the power in the other phantom loop antenna is left unaffected. By varying the ratio of power in the two phantom loop antennas by means of resistance the courses may be made to shift from their normal 90-deg. displacement to any angle between 45 deg. and 135 deg. See Fig. 5. It should be carefully noted that this variation depends upon several variables including loop antenna resistance, mutual induction between primary and secondary of goniometer, and power factor of link circuits. These should be taken care of jointly in each particular installation.

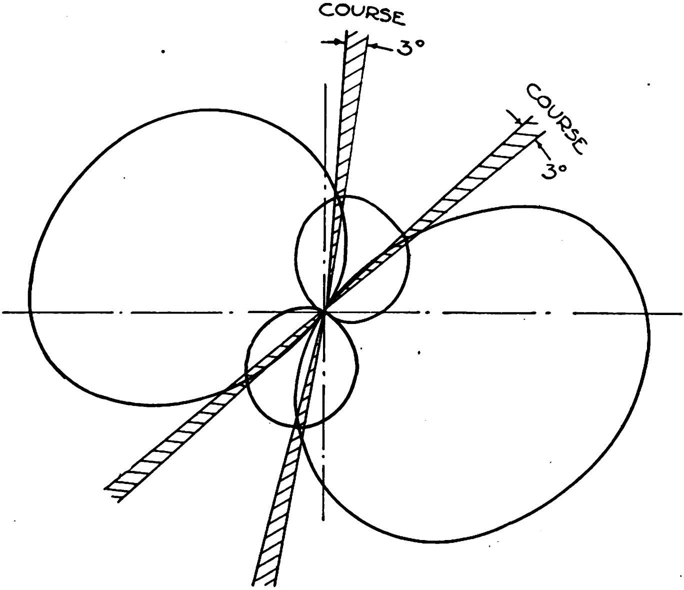


Fig. 5—Observed field-intensity patterns of each phantom loop antenna with 32 ohms resistance inserted in series with one goniometer primary; goniometer primaries set at 90 deg.

Method Used for Courses Emanating at Angles between 135 Deg. and 180 Deg. The normal polar pattern of crossed loop antennas is a pair of figures-of-eight at 90 deg. This gives four equisignal zones which are referred to as "courses" in beacon terminology. See Fig. 6. In order to increase the sharpness of the course the goniometer primaries are frequently set at 120 deg. instead of 90 deg. This gives four courses, two of which are about 3 deg. wide, while the remaining two are 12 deg. wide. See Fig. 4.

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[2] A phantom loop antenna is the resultant figure-of-eight space pattern occurring when both loop antennas are coupled to one goniometer primary. The angular relation of the phantom loop antennas is the same as that of the goniometer primaries. For a further treatment of this subject see Murphy and Wolfe, "Stationary and rotating equisignal beacons," *Jour.* S. A. E., 19, 209; Sept., 1926.

If a vertical wire is now permitted to radiate in time phase with the radiation from the phantom loop antenna the field-intensity pattern will be distorted as shown in Figs. 7 and 8. It will be noted that two lobes diminish and two increase. This changes the relative position of the equisignal zones so that the course approaching the beacon is no longer 180 deg. from the one leaving. The amount of this shift may be controlled by adjusting the magnitude of the current in the vertical wire. This shift is limited only by the disappearance of two of the lobes,[3] forming two cardioid patterns at right angles, thus giving two equisignal zones displaced 180 deg. instead of four displaced at various angles.

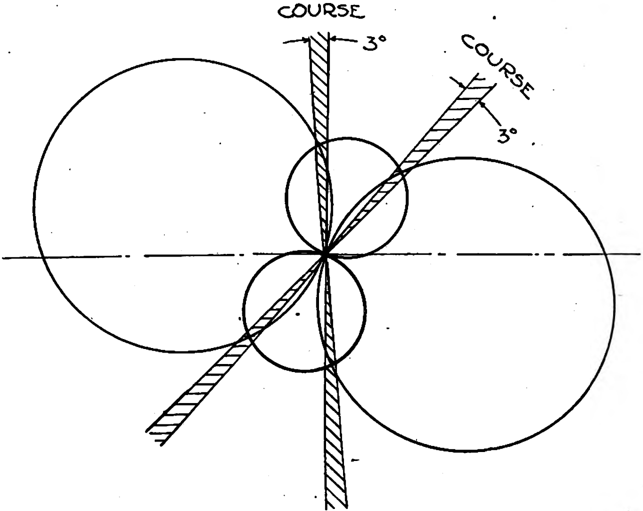

Fig. 5a—Theoretical field-intensity patterns of each phantom loop antenna with 32 ohms resistance inserted in series with one goniometer primary; goniometer primaries set at 90 deg.

It has previously been mentioned that in the original beacon system the goniometer primaries were set at an angle of 120 deg. to secure a narrow course. When means are employed to bend the course, this is not advisable. There is danger of one lobe shrinking so far that it intersects the adjacent lobe on the wrong portion of its contour. This will be apparent from Fig. 9. Four courses are present so far as the ear can detect, but as the lobe decreases the course will no longer

[3] E. Z. Stowell, "A unidirectional radio beacon for aircraft," Bureau of Standards Journal of Research, December, 1928.

shift properly. Furthermore, the course at this intersection is not sharp enough to be used. In order to prevent this in beacon installations the primaries are frequently set at 90 deg. when the courses are to be bent.

III. Application of Theory

Having decided upon a method for aligning the courses of the radio range with the airways, the next problem was to develop a means of applying it to the beacon installations. The radio ranges at Hadley Field, New Brunswick, N. J., and Bellefonte, Pa., were selected for these tests.

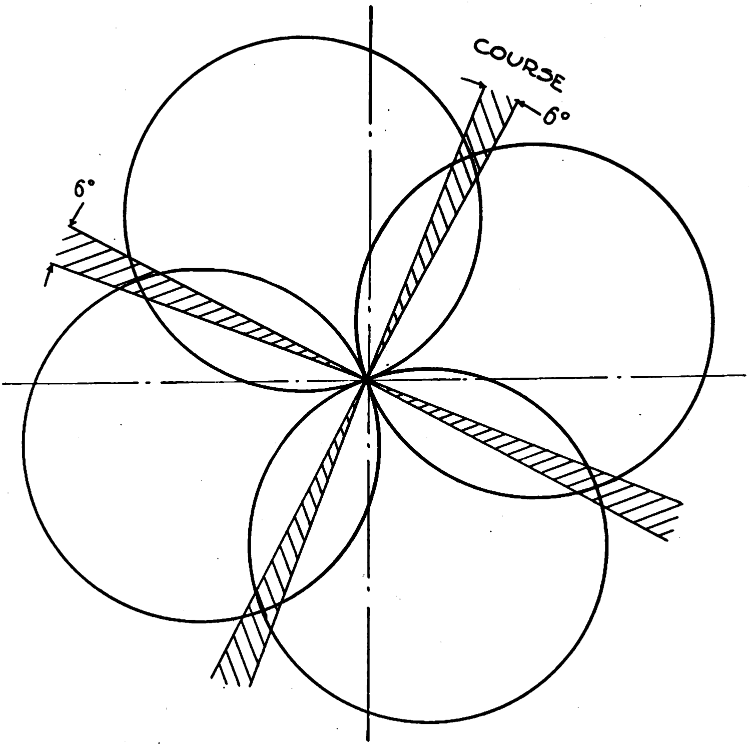

Fig. 6—Theoretical field-intensity patterns of each phantom loop antenna energized separately; goniometer primaries set at 90 deg.

The airway route through Bellefonte has a 14-deg. bend at Bellefonte, and consequently, the antenna effect method was employed.

A vertical wire antenna was installed on the radio range tower and this was tuned to resonance by a suitable series inductance. The circuit was of sufficiently high resistance to have a broad resonance curve.

The antenna was at first excited by coupling it to the master oscillator. The results from this were poor, and analysis showed that a pronounced phase shift occurred between the master oscillator and the goniometer, due to the particular coupling circuits employed. To eliminate this, a coupling coil was inserted in the common goniometer primary circuit, and the antenna coupled to this. This resulted in excellent control of the amount of shift, displacements of 28 deg. being readily obtainable. The goniometer primaries were set at 90 deg. for this work.

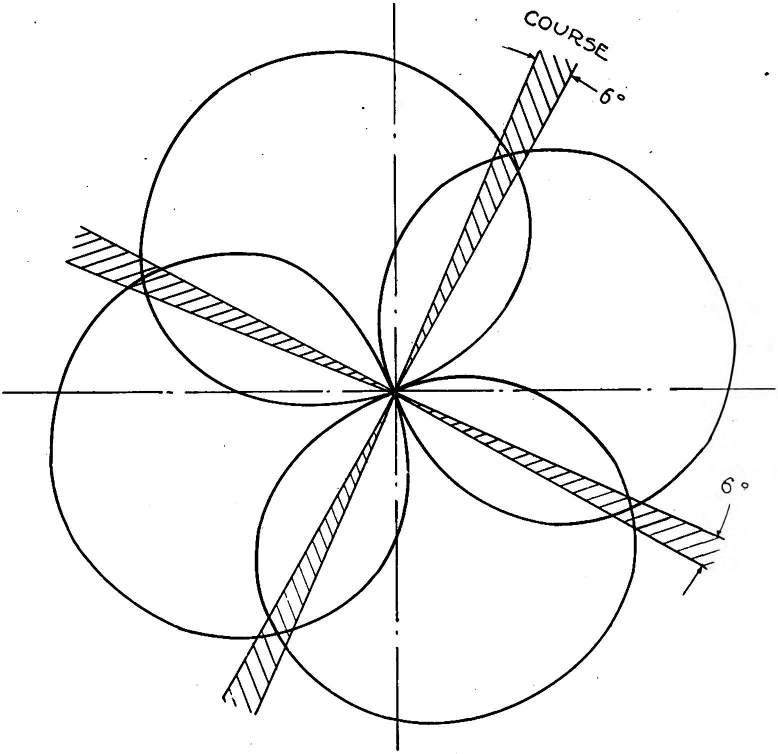

Fig. 6a—Observed field-intensity patterns of each phantom loop antenna energized separately; goniometer primaries set at 90 deg.

The final adjustment of antenna current was made by controlling the degree of the antenna coupling. It is necessary that the antenna be in tune to prevent dephasing; consequently, detuning will not accomplish the desired result.

It was also necessary to have an accurate knowledge of the amount of shift, in order that the course might be properly oriented. For

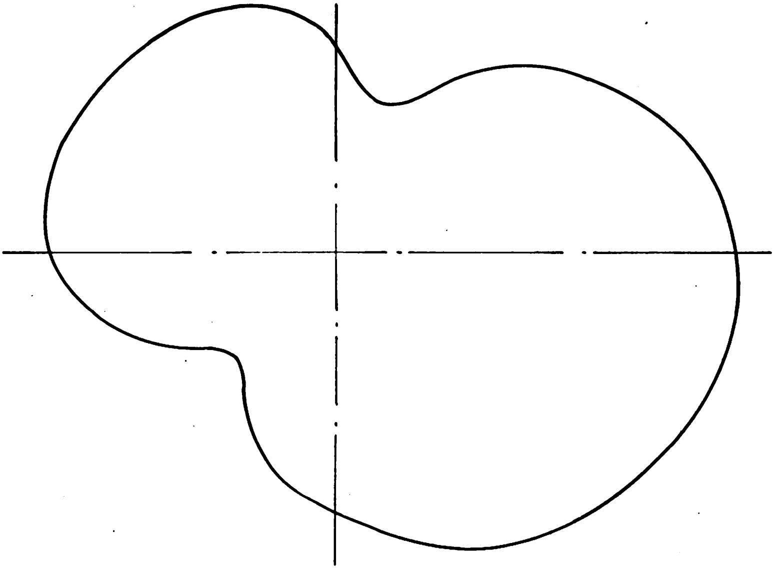

Fig. 7—Observed field-intensity pattern of one phantom loop antenna and vertical antenna energized simultaneously.

first adjustments a receiving set using a six-foot vertical antenna was mounted in a truck and driven around the loops at a distance of about 100 yards from the base of the tower. Readings were very sharp at

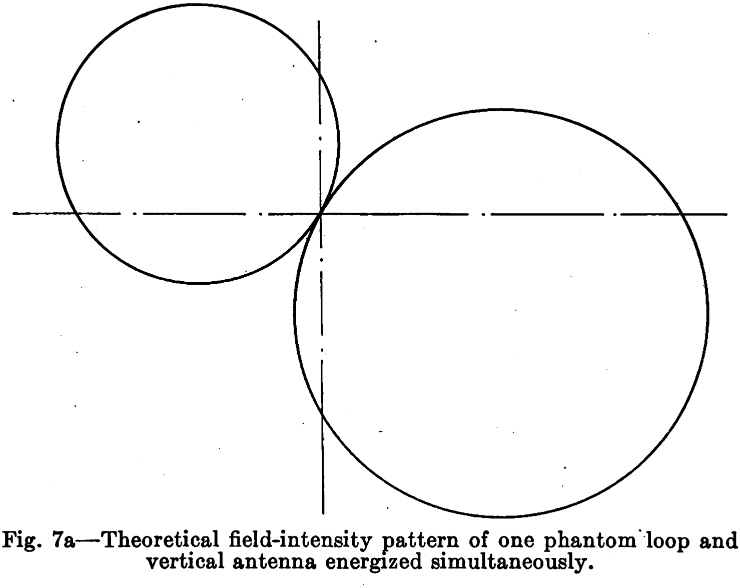

Fig. 7a—Theoretical field-intensity pattern of one phantom loop and vertical antenna energized simultaneously.

this distance and the 14-deg. bend was easily checked. By locating landmarks from compass bearings it was possible to align the course with the airways.

In order to secure the observed field-intensity patterns shown, the same receiving set was employed with its vertical pole antenna. The output was connected to a calibrated thermoelement and the signal intensity adjusted to keep the tubes under saturation at all times. In employing the vertical wire antenna for distorting the pattern the receiving set was stationary and the goniometer rotated through 360 deg. With the series resistance method, the set was moved around the beacon tower.

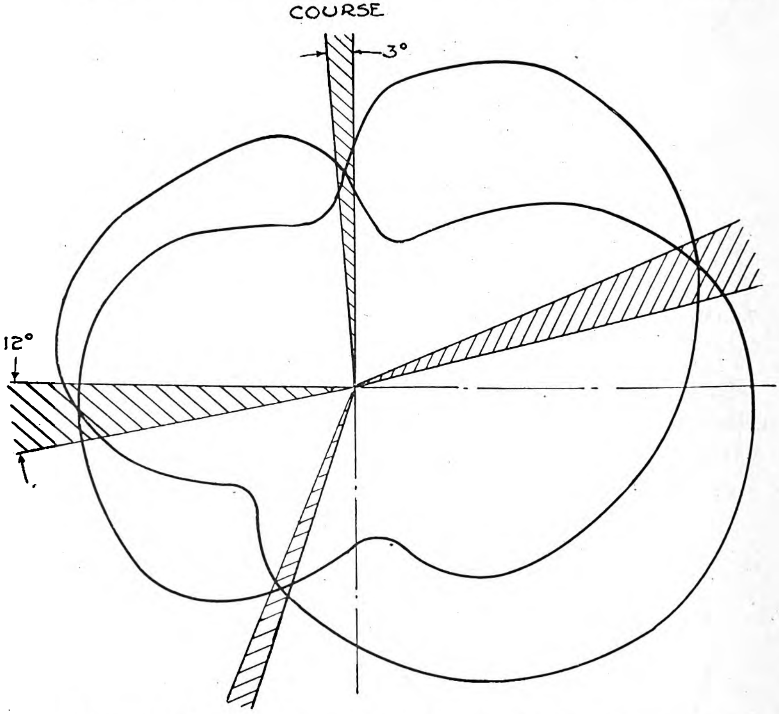

Fig. 8—Observed field-intensity patterns of each phantom loop antenna and vertical antenna energized separately; goniometer primaries set at 60 deg.

Discrepancies between observed and theoretical patterns are due chiefly to phase shift between the loop antenna currents and the vertical wire antenna current.

For a further check the truck was taken to Snowshoe Junction to the west and to Aaronsburg to the east. (These are, respectively, ten and twenty miles from the radio range). The readings here checked with those taken at the beacon tower to within half a degree, which is well within the accuracy of measurement. As a final check, the course was flown again, and once more the readings checked those taken at the beacon. From this and from data secured on later occasions, it was decided that an accuracy of 1 deg. could be expected from measurements made at the beacon tower, in spite of the predominating induction field.

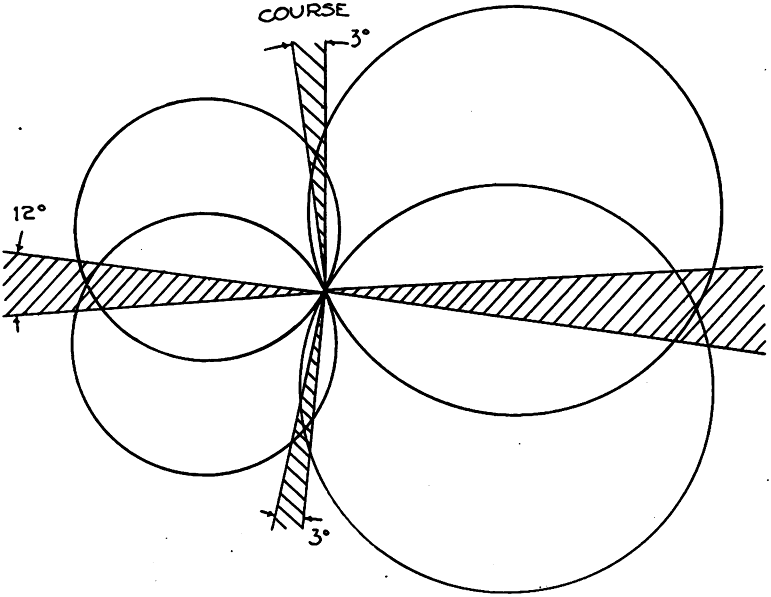

Fig. 8a—Theoretical field-intensity patterns of each phantom loop antenna and vertical antenna energized separately; goniometer primaries set at 60 deg.

Hadley Field, besides being the eastern terminus of the Transcontinental Airway, has two other air lines entering it. Hartford to the north, and Washington to the south, need to be served by the same radio range. Washington and Hartford lie in a straight line through Hadley Field, but the course to the west is not at 90 deg. to these. This presented an opportunity to use the first method described, namely that of inserting resistance in the link circuit of one of the goniometer primaries, which, as before explained, shifts the 90-deg. courses.

With the current in one goniometer primary approximately one-half the value of the other, the courses were shifted the desired amount.

Here also it was found desirable to adjust the goniometer primaries to 90 deg. in order that the courses might be of equal power. This change in relative field intensity of the two phantom loops has another advantageous effect in that it sharpens the course to a considerable degree. By referring to Fig. 5 it will be seen that since the intersection of the smaller figure-of-eight comes at the point of rapid change of the large figure-of-eight, the resulting course is quite sharp. By measurement the courses have been found to be approximately 3 deg. wide. This is much sharper than the usual course resulting from a 90-deg. separation of the goniometer primaries.

The courses were checked here in a manner similar to that used at Bellefonte.

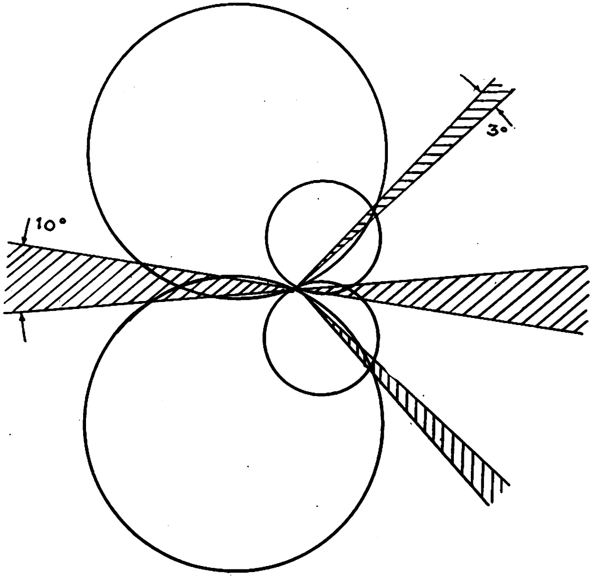

Fig. 9—Result of incorrect time phase of vertical antenna.

Another modification of the system is possible. When using a vertical wire antenna it may be keyed by a relay in the same circuit as the one goniometer primary. The result of this will be to affect only one phantom loop and thus shift both the 90-deg. and the 180-deg. courses.

With these three means at hand it becomes a relatively simple matter to satisfy almost any requirement where the number of courses to be served does not exceed three, and also a great number of four-course air terminals. The adjustments are not so critical as to be unstable, and once made remain indefinitely. Very little extra apparatus is required and the maintenance is not difficult. All the requirements for the aural type of radio range, as listed in the early part of this paper, are therefore fulfilled.

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EFFECT OF FLIGHT ON HEARING*

BY
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Summary—*The unusual conditions imposed by flight on the sense of hearing are stated, showing the need of audiometric study. After brief review of the theory of the audiometer, data are submitted covering:* (*a*) *fatigue as indicated by measurements before and after flight,* (*b*) *fatigue as indicated by measurements taken during flight,* (*c*) *initial impairment coincident with flight,* (*d*) *group measurements of fliers and non-fliers as a test for permanent impairment. The results are discussed with particular view to their bearing on aircraft radio operation.*

OF the senses engaged in flying probably none is subjected to more unusual conditions than that of hearing. The ears receive during flight a continuous and intense bombardment of complex sound caused by the motor exhaust, propeller, and wind noises. There are also changes of pressure and temperature of greater range and rapidity than are usually encountered elsewhere. Because of these conditions, conversation can be carried on in a closed cabin plane with greatest difficulty only, while in an open cockpit during normal flight it becomes utterly impossible.

This condition was early recognized as a dominant factor in the use of radio in airplanes, and it was at first doubted if such use was possible except with visible means of reception. However, by employing special telephones and protective covering to the ears, remarkable success has been attained in spite of the highly adverse conditions.

In addition to the initial reduction in efficiency of the flight radio operator due to flight conditions, there is a cumulative effect due to fatigue as evidenced by the temporary deafness of fliers noticeable immediately after flight. It was suggested after the memorable transatlantic flight of Commander Byrd that one reason for the ineffectiveness of the radio during the last critical hours of this flight was the result of this fatigue over a long period.

As this subject appeared to be one of immediate practical importance, as well as of some scientific interest, an investigation was undertaken by the United States Naval Research Laboratory beginning in July, 1927. This work was authorized by the Bureau of Medicine and Surgery, and carried out at the Anacostia Naval Air Station.

The tool for making the study was available in the form of a standard Western Electric audiometer. This instrument is based on re-

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* Dewey decimal classification: R529. Original manuscript received by the Institute, October 28, 1929.

search work of the Bell Telephone Laboratories which is covered in a comprehensive paper by Harvey Fletcher.[1] In order to make the results here presented more easily interpreted by those unfamiliar with this instrument, a brief outline of the underlying theory is offered.

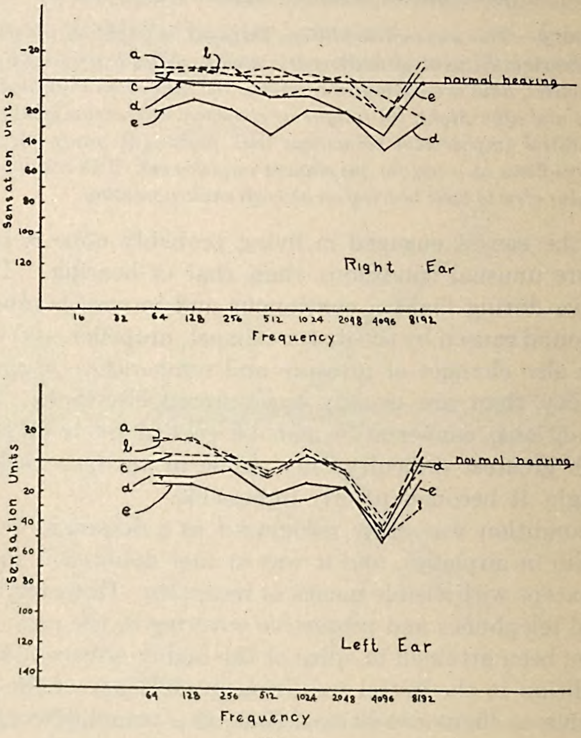

Fig. 1

(a) 11/7/27; before.
(b) 11/8/27; before.
(c) 11/9/27; before.
(d) 11/9/27; after. Time, 40 min. Plane, D. H. Motor, Liberty. Position, front cockpit. Altitude max., 3000. Altitude avg., 2500. Motor speed, 1400.
(e) 11/9/27; after. Time, 2 hr. 20 min. Plane, D. H. Motor, Liberty. Position, front cockpit. Altitude max., 8000. Altitude avg., 7500. Motor speed, 1300.

A pure sound has two qualities or dimensions which lend themselves readily to quantitative measurement, namely amplitude and frequency. This suggests a plain figure or area to represent such measurements. Experiment determines the limit of frequency to which the normal ear will respond as from 16 to 16,000 cycles per second. Likewise the maximum limits of intensity which lie in the mid frequency region, when measured in dynes r.m.s. pressure on the ear drum, fall between 0.001 and 10,000 dynes. This fixes the limiting dimensions of the area of hearing.

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[1] *Jour.* Franklin Institute, 196, 289, 1923.

The scale for such dimensions is next determined. From observation of the least discernible increment which the ear can detect, it develops that such increments are not constant, but increase logarithmically both for frequency and intensity. It is convenient to choose for the intensity scale a unit which is twenty times the common logarithm of the approximate ratio of least discernible increments in the r.m.s. pressure on the ear drum. This unit is the same as the decibel. For the frequency scale the logarithm of the one hundredth root of two ($\sqrt[100]{2}$) is chosen. The area of audible sound plotted in this fashion has a maximum amplitude dimension at 1000

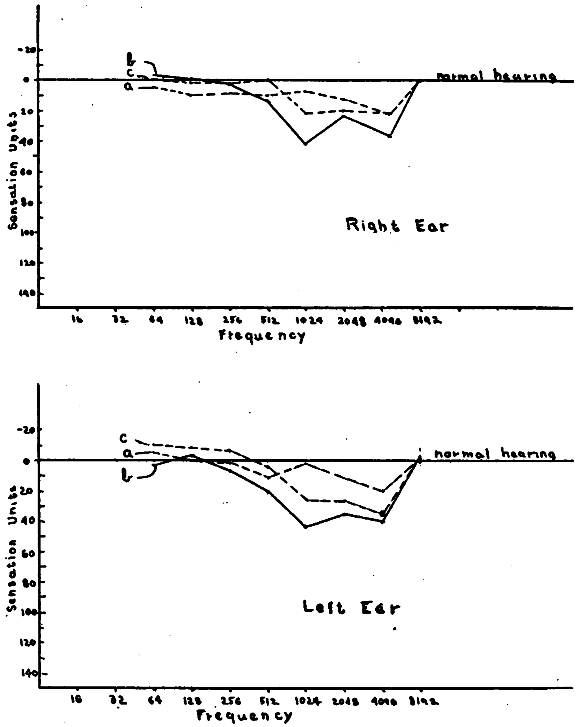

Fig. 2

(a) 11/14/27; before.

(b) 11/16/27; after. Time, 1 hr. 50 min. Plane, 02B2. Motor, Liberty. Position in plane, rear cockpit. Altitude max., 1600. Altitude avg., 1500. Motor speed, 1500 R P M.

(c) 11/16/27; before.

cycles per second of 270 increments, and a maximum frequency dimension near the one dyne pressure level of 1300 increments.

The entire area represents a total of some 300,000 sounds or sensation units, each differing from all the others in either frequency or intensity. This is the sensation unit which is the basis of the audiometric chart used with the audiometer.

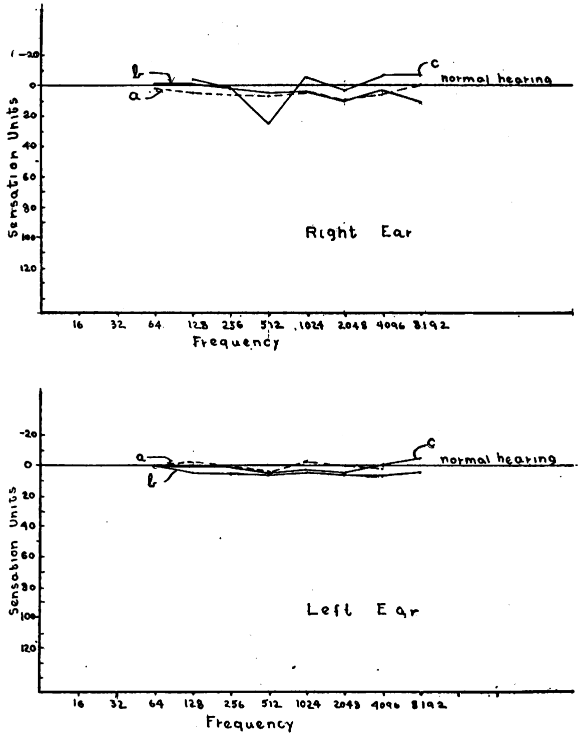

Fig. 3

(a) 9/13/28; before.

(b) 9/13/28; after. Time, 1 hr. 15 min. Plane, V01. Motor, Whirlwind (J5). Position in plane, rear cockpit. Altitude avg., 4500. Altitude max., 5000. Motor speed, 1700.

(c) 7/18/29; after. Time, 7 hr. 20 min. Plane, T4M1. Motor, Pratt Whitney. Position in plane, radiocompass. Altitude max., 5000. Altitude avg., 3000.

The audiometer itself consists of a vacuum-tube oscillator for producing sounds of different frequency, an attenuator for reducing such sounds, and a telephone through which they are applied to one ear of the person whose hearing is to be investigated. The frequency at which the instrument operates is not continuously variable, but changes by multiples of two from 32 cycles to 8192 cycles per second. The attenuator scale, modified for each frequency, reads directly in sensation units. In use, the attenuator is set for the weakest sound which the subject can hear. The scale then reads loss of hearing in sensation units at that frequency. Dividing by the total number of sensation units discernible to the normal ear at said frequency gives the loss of hearing in percentage.

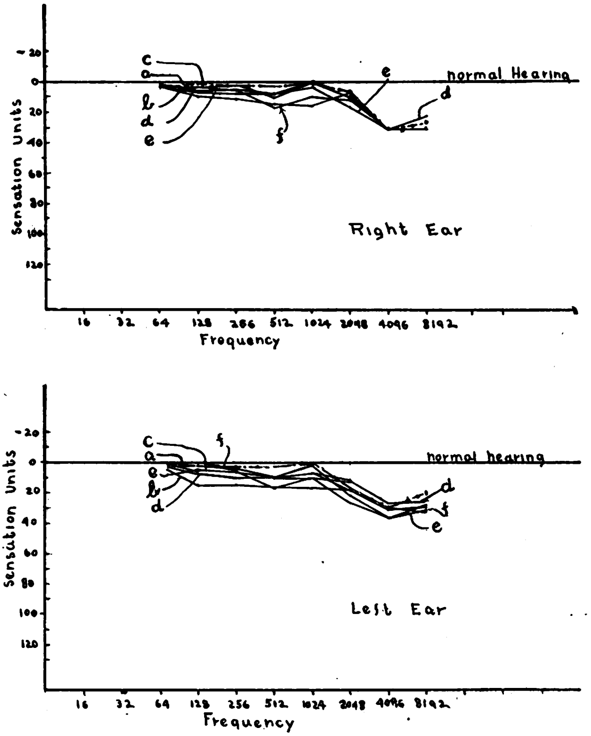

Fig. 4

(a) Average, before 9/12/28; 9/14/28; 9/25/28; 9/28/28.
(b) 9/10/28; after.
(c) 9/14/28; after. Time, 1 hr. 30 min. Plane, V01. Motor, Whirlwind-J5. Position, rear cockpit. Altitude max., 5000. Altitude avg., 5000. Motor speed, 1800.
(d) 9/25/28; after. Time, 50 min. Plane V01. Motor, Whirlwind-J5. Position, rear cockpit. Altitude max., 5000. Altitude avg., 3000. Motor speed, 1750.

It was first undertaken to determine the effect of fatigue by taking measurements before and after flying. The measurements before flight were repeated on several days to check accuracy of procedure and magnitude of variations due to causes other than flight fatigue. The results of such preliminary measurements are shown in separate dotted lines of Figs. 1 to 6, except that to avoid confusion the dotted line in Fig. 4 is the average of four separate measurements which fell close together.

Each of the subjects for Figs. 1, 2, 3, and 4 had flown for a period of years with total flight time approaching or in excess of 1000 hours, and each has fairly normal hearing. Case one is for a pilot, while the other three are aircraft radio operators of exceptional experience and ability. Figs. 5 and 6 are given to show effect of flight where hearing is abnormal. These cases were for observers who had done intermittent flying for a period of years, but with less total flight time than any of the first four. There is no direct evidence that the defects of hearing in either of these cases are in any way due to flying. Fig. 7 shows the results of measurements taken during actual flight to determine the initial loss of hearing due to flight conditions as well as the cumulative effect. Fig. 9 is the result of measurements of two groups, one consisting of 22 men whose duties require regular flights of at least four hours

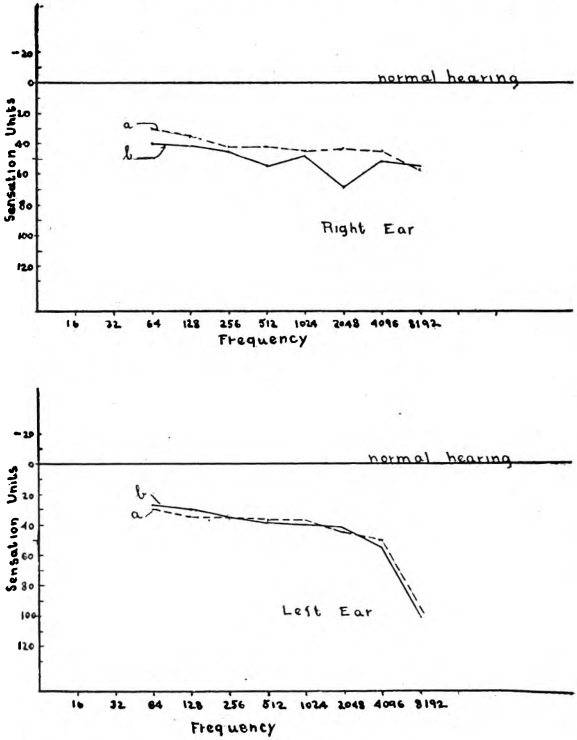

Fig. 5

(a) 9/17/28; before.
(b) 9/17/28; after. Time, 45 min. Plane, V01. Motor, Whirlwind. Position, rear cockpit. Altitude max., 5000. Altitude avg., 5000.

per month, while the other includes 45 men whose duties require no flying. They have, therefore, flown only occasionally if at all. The purpose of this part of the survey was to determine whether or not flying produced any permanent effect on hearing.

In discussing the results, it may first of all be noted that the method of measurement before and after flight, as shown in Figs. 1, 2, 3 and 4, gives definite evidence of fatigue, though it varies widely with the individual. On the other hand, method of measurement at successive periods during a long continued flight, as shown in Figs. 7 and 8, shows no evidence of fatigue. This might seem a direct conflict. However, in the first case the measurements were made on a single pure tone in a quiet room, while in the second case the measurements are at a high intensity level, in the presence of strong interference or masking. Putting it another way, the ground measurements deal with only the lower edge of the hearing area previously referred to,

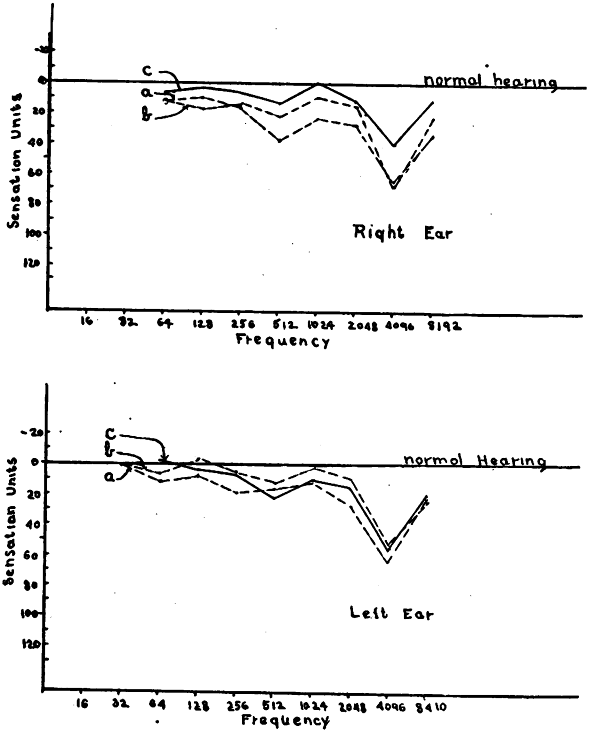

Fig. 6

(a) 9/29/27; before.
(b) 9/30/27; before.
(c) 9/30/27; after. Time, 30 min. Altitude, 8000.

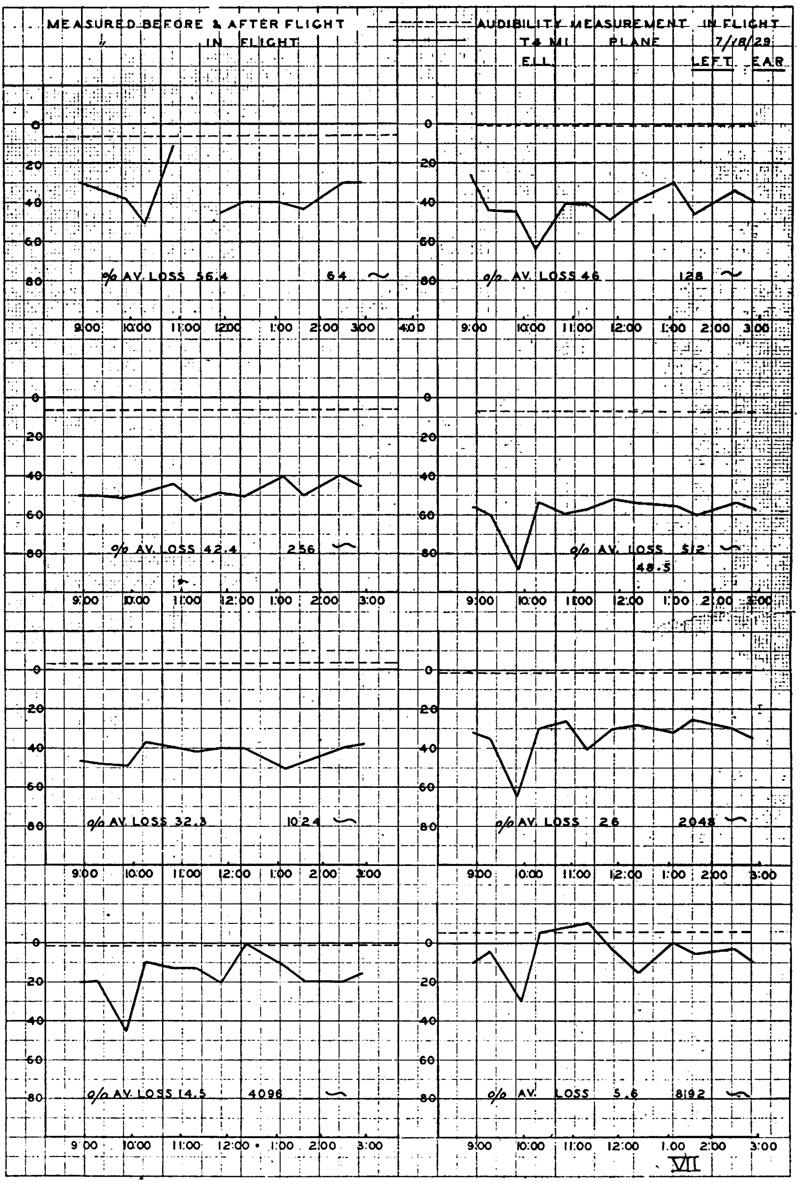
MEASURED BEFORE & AFTER FLIGHT
" IN FLIGHT
AUDIBILITY MEASUREMENT IN FLIGHT
T4 MI PLANE 7/18/29
E.L.L. LEFT EAR
0
20
40
60
80
% AV. LOSS 56.4
64 ~
9:00
10:00
11:00
12:00
1:00
2:00
3:00
4:00
% AV. LOSS 46
128 ~
% AV. LOSS 42.4
256 ~
% AV. LOSS 512 ~
48.5
% AV. LOSS 32.3
1024 ~
% AV. LOSS 26
2048 ~
% AV. LOSS 14.5
4096 ~
% AV. LOSS 5.6
8192 ~
VII

Fig. 7

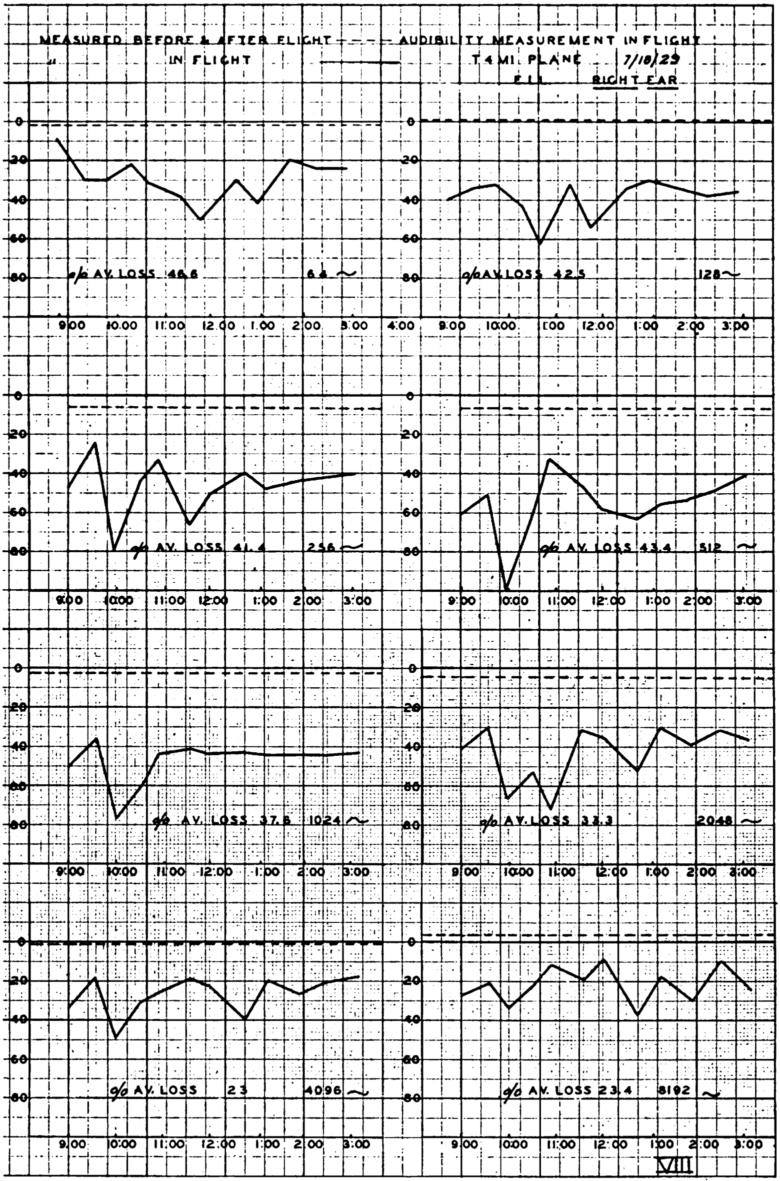

Fig. 8

22 Fliers———45 Non Fliers

Hearing Loss Per cent					*Frequencies*				
Right ear	64	128	256	512	1024	2048	4096	8192	Avg.
(a) Fliers	8.6	6.4	8.85	11.7	9.6	12.15	26.2	22.	13.19
(b) Non-fliers	10.9	10.	11.7	11.9	7.	8.5	17.05	15.7	11.59
								diff.	1.60
Left Ear									
(a) Fliers	5.9	7.	7.75	10.8	8.6	14.2	25.5	21.	12.59
(b) Non-fliers	12.9	8.7	9.8	11.35	7.	9.5	16.15	12.95	11.04
								diff.	1.55

while the flight measurements deal with a region well up toward the middle. There is evidently, then, a wide difference between the conditions under which the measurements are taken, and the conception that fatigue might be evident in one case and not in the other is not untenable.

It is the second set of measurements, however, that are of the most practical importance from the standpoint of aircraft radio comunication, as these represent the actual conditions under which the operator works; therefore, the evidence that the subject of this test had lost none of his ability to hear signals after eight hours continuous flight is considered of great interest. Again the personal equation must be considered. The subject for this test was the same as for test of Fig. 4, and the record in that case also shows no fatigue, except the loss at 500 cycles shown on curve (c), which is doubted, as there may have been outside interference when this point was taken.

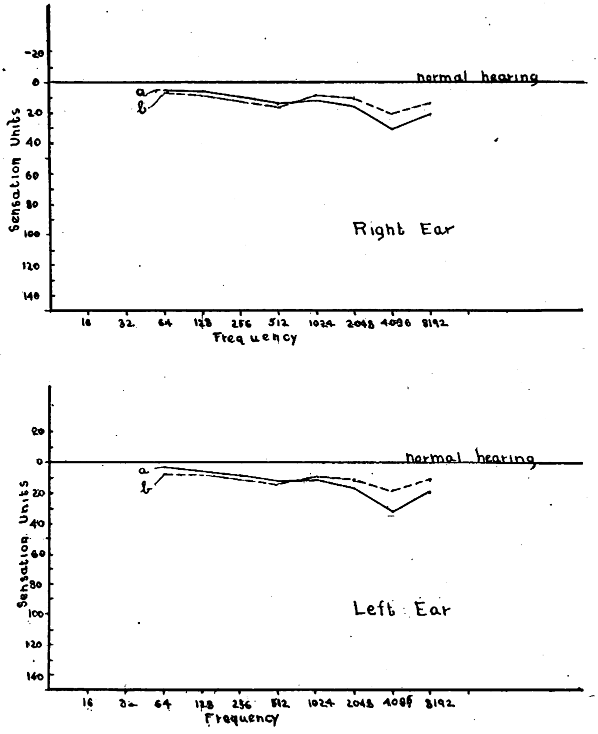

Fig. 9

Another condition bearing on this case is that during the flight measurements the subject was in the radio space of a bombing plane and was much better protected against flight noises than had he been in an open cockpit, as were the subjects during most of the other measurements. But these are the conditions which the radio operator on a long flight would be most apt to have.

Weighing all the evidence here presented, it seems doubtful that flight noises alone will produce any cumulative action which would effect the efficiency of a radio operator during even the longest flights. However, there is still a possibility that such a cumulative effect may result indirectly from the abnormally loud signals required in the telephones to get through the masking of flight noises. It has been noted that a voltage of ½ to 1 volt at telephones is required to be audible in flight, and that during flight such a signal is *barely* audible, while a signal of equal strength on the ground seems almost intolerably loud to the normal ear. One of the experienced operators used as a subject during these tests remarked that he did not experience the buzzing sensations and deafness after flight that he had previously noticed after flights during which he had handled radio traffic almost continuously. In addition to outside signals, his set would naturally be adjusted so that he could hear himself send on his own transmitter, which was of a type confined to output on a single audible frequency. It is then suggested that continued exposure to still more intense sound confined to a single frequency may produce serious cumulative fatigue, though a complex sound covering much of the sound spectrum does not. This point is not covered by experimental work here reported, and therefore remains for future investigation.

It is further to be noted that there is no evidence in Figs. 1 to 4 that fatigue is noticeably intensified at any particular frequency. Neither is one conscious of any excessively pronounced pitch when listening to the flight noises. Though the audiometer does not permit measurement at all frequencies, but only at multiples of one particular frequency, it explores enough of the total area of audition so that the possibility that fatigue at any particular frequency could have existed without detection seems remote.

Figs. 5 and 6 were taken to show the effect of flight on an abnormal ear. Fig. 5 is in line with those preceding it in that such change as is shown in the measurements taken after flight is in the nature of a loss of hearing. Fig. 6, however, is unique in that the right ear showed a marked *gain* after a short flight at high altitude, though the gain is not apparent in measurements of the left ear. This subject is known to have a catarrhal condition of his right ear such as would very possibly produce stoppage of the eustachian tube. It is suggested that the rapid pressure variation due to the quick change in altitude may have cleared up this stoppage in much the same way that a physician would have done by catheterizing the ear, which might well account for the gain indicated. This case may throw some light on the claims sometimes advanced in the press of "cures" of deafness produced by flying.

Passing to the question of initial impairment of hearing under flight conditions, irrespective of fatigue, Figs. 7 and 8 yield interesting information. The subject for this test was chosen because of his unusually normal hearing, which averaged for all frequencies practically 100 per cent. This is indicated on the charts for each frequency by the dotted lines. The loss in sensation units during flight is shown by the solid lines, while the average percentage loss of hearing is given in Table I.

TABLE I

AVERAGE PERCENTAGE LOSS OF HEARING DURING FLIGHT FROM FIGURES 7 AND 8

Frequency		*Right Ear*	*Left Ear*
64		46.6	56.4
128		42.5	46.
256		41.4	42.4
512		43.4	48.5
1024		37.6	32.3
2048		33.3	26.
4096		23.	14.5
8192		23.4	5.6
	Average	36.4	33.9
	Average both ears,		35.1

This shows that a normal operator during flight is equivalent in aural efficiency to a man on the ground who has lost about one third of his hearing. That is, one third the total number of sounds making up the sensation area for the normal ear are inaudible to him. It further indicates that the loss diminishes as the frequency increases. This is in line with some of the work on masking.[2] It immediately suggests itself that a gain might result if radio communication in flight were carried on at a high audible frequency. However, the normal ear is most sensitive to pitches around 1000 cycles per second, and falls off for frequencies both above and below this point.

Therefore, though the loss percentage at higher frequency may be less, the efficiency at lower frequency may remain greater. In any event code reception in planes is usually by continuous wave where the pitch of the received note is under the ready control of the operator. Therefore, he naturally makes the selection of pitch which best suits the conditions and his individual ear. For telephone communication, of course, no selection of audible frequency is possible.

Considerable irregularity in measurements is indicated in Figs. 7 and 8, particularly during the early part of the record. This may have been due to variation of flight noises with change of engine speed and other conditions of flight, or it may have resulted from lack of experience in taking audiometric measurements under flight conditions. As the subject became more familiar with the procedure he gave a more uniform and positive response to the sound of the instrument. It is not believed that these variations could affect the conclusions which have been drawn from this part of the work.

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[2] H. L. Wegal and C. E. Lane, "Auditory Masking and Dynamics of the Inner Ear," *Phys. Rev.*, February 1924; *Proc.*, American Physical Society, April, 1923.

An angle of the study of interest to all fliers, radio operators or not, is brought out in Fig. 9. This deals with the question of whether the hearing of aviators is permanently affected by continued flight. It shows the measurement of thirty-two fliers as compared with forty-five non-fliers. Many of the measurements were repeated one or more times, so that about 150 sets were taken in all. Unfortunately, no detailed information is available as to the total individual flight time of the flying group, but they were most all experienced men who had flown regularly and continuously for a period of years. The results show an average loss by the fliers about 1.5 per cent greater than for the non-fliers, most of this being at the higher frequencies. Thus at 4096 cycles the hearing of the fliers falls below that of non-fliers about 8 per cent. Whether this lowering of sensitivityat the upper frequencies is a particular effect of flying, or because the general run of ear troubles involve these frequencies first, is not determined. While there is this evidence of a definite permanent loss of hearing resulting from flight, it is quite clear that it does not always result. For example, the subject of Figs. 4, 7, and 8, who, as previously stated, shows exceptionally good hearing, is one of those who has done the most flying.

While it is felt that the subject is far from exhausted, and that the work here described is by no means complete, fairly definite conclusions on several important points seem to be justified:

(a) From the standpoint of the normal radio operator on the ground the best operator in flight is at least one third deaf and his receiving equipment should be designed with due reference to this fact.

(b) While there is definite evidence of fatigue *after* flight, the fatigue produced by normal flight noises does not appear to be selective as to frequency.

(c) There is no evidence of fatigue produced by normal flight noises *during* flight.

(d) There is evidence of a small but definite permanent loss of hearing among experienced aviators, though such loss is not universal and is limited to the high frequencies.

(e) It would seem highly desirable that service men chosen for training as aircraft radio operators be carefully selected by means of audiometric measurements; to be sure that they possessed good hearing, particularly at the higher frequencies. Moreover, such measurements should be repeated from time to time as their service progressed.

It is desired to acknowledge the cooperation of the officers and men of the Anacostia Naval Air Station who became subjects for these tests, and to express particular appreciation of the collaboration of the medical officer, Lt. Comdr. W. W. Davies, under whose supervision the work was carried out.

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BOOK REVIEWS

Radiotechnik: VI. Die Elektrischen Wellen, by DR. F. KIEBITZ. Sammlung Göschen, Berlin and Leipzig, Walter de Gruyter & Co., 1929, 125 pages. Price 1.50 RM.

This excellent semi-popular account of electric waves forms the sixth volume on radio to appear in the well-known Göschen series. Commencing with the elements of electrical theory, the book deals successively with the Hertz experiments, radiation resistance, the nature of electromagnetic waves, and wave propagation around the earth. Considerable space is devoted to directive antennas. Unfortunately but little attention is paid to the ionized layer, skip distances, and echoes.

W. G. CADY†

† Wesleyan University, Middletown, Conn.

Elements of Radio Communication, by JOHN H. MORECROFT, John Wiley & Sons, Inc., 1929. 269 pp. 170 illus.

"The Elements of Radio Communication" is an adequate and comprehensive text book on radio communication stressing particularly the field of broadcast receiver design. It develops the subject in a natural and conventional progression using language suited to the needs of an elementary college course. The fundamental principles of apparatus design are largely applied to the broadcast receiver and its associated equipment. The mathematical treatment, with very few exceptions, does not range beyond the field of advanced algebra.

As one would expect from such an experienced teacher as Professor Morecroft, his scholarly style and workmanlike presentation have been skilfully applied with the requirements of instruction purposes in mind. Practical problems for each chapter are given to test the student's mastery of the text. The keynote throughout the work is practical application rather than theoretical dissertation. Complete data descriptive of the vacuum tubes used in modern receivers, radio-frequency amplifiers of conventional design used in broadcast reception, and the full gamut of associated equipment, such as power supply devices and reproducers, are adequately covered. Although broadcast transmission equipment is naturally stressed in the consideration of that field, the spark transmitter, the high-frequency alternator, and single sideband transmission are not overlooked.

The work is a needed contribution to radio literature, successfully bridging the gap between the more comprehensive studies of the subject, of which there is no better example than the same author's "The Principles of Radio Communication," and the "popular" works on the subject which confine themselves to the generalities. It is suited to the needs of the embryo commercial engineer and laboratory worker and as a text book for elementary college courses.

EDGAR H. FELIX‡

‡ Radio consultant, National Electric Manufacturers' Association, New York City.

BOOKS RECEIVED

Transmission Networks and Wave Filters, by T. E. Shea, member of the technical staff, Bell Telephone Laboratories. Published by D. Van Nostrand Company, Inc., 1929. 470 pages, illustrated, 6 x 9 inches, cloth binding. Price $6.50. Contents: Part I, principles of transmission networks; external performance of networks—conditions for network equivalences—some important network parameters; effect of impedance mismatching at the terminals of a network—reflection factor—interaction factor; properties of two-terminal impedance arms. Part II, electric wave filters: general—conditions for free transmission and for attenuation; the more common types of filter sections; the formation of multisection filters—low-pass, high-pass, band-pass, and band-elimination types; terminal effects: reflection and interaction losses—total insertion loss characteristics. Part III, composition of transient waves: resolution of steady state waves into frequency components by Fournier series analysis; resolution of transient waves into frequency components by Fournier integral analysis. Appendix A, bibliography; Appendix B, relation between decibels and current, voltage and power ratios; index.

Fundamentals of Radio, by R. R. Ramsey, Professor of Physics, Indiana University. Published by Ramsey Publishing Company, Bloomington, Indiana. 372 pages, illustrated, 6 x 9 inches, cloth binding. Contents: electricity, direct current; batteries; measurement of resistance; alternating current; introduction to radio; capacity; inductance; radio waves—radio current—transmission; detectors; vacuum tubes; vacuum-tube constants; the vacuum tube used as a detector; the tube as an amplifier; the tube as an oscillator; coupled circuits; how radio messages are transmitted by the ether; radiation from an aerial; aerials; radio-frequency instruments and apparatus; radio resistance; spark transmission; long-wave CW transmitters; vacuum-tube transmitters; radio telephone; receivers; audio amplification; balanced circuits; loud speakers; rectified alternating current; applications of the vacuum tube.

BOOKLETS, BULLETINS, AND CATALOGS RECEIVED

The Central Scientific Company of 460 East Ohio Street, Chicago, Ill., has available for free distribution copies of its new technical publication, "High Vacuum Engineering," which deals with vacuum pumps and their specific application to the exhaustion of radio tubes.

The "Resistance Handbook" of the Gilby Wire Company, Newark, N. J., contains a considerable amount of data concerning nickel and its alloys, particularly those used in radio equipment.

Two booklets, "About Vitreosil" and "Vitreosil for the Laboratory" are available from the Thermal Syndicate, Ltd., 58 Schenectady Avenue, Brooklyn, N. Y. These describe the uses of, its characteristics, and the forms in which vitreosil may be obtained.

"Rare Metals," distributed by the Fansteel Products Co., Inc., of North Chicago, Ill., is a booklet devoted to the history, properties, and uses of tantalum, tungsten, and molybdenum.

A booklet describing bakelite varnish can be obtained from the Bakelite Corporation, 247 Park Avenue, New York City.

A very brief list of the physical and electrical characteristics of isolantite may be obtained from the Isolantite Company of America, Inc., 551 Fifth Avenue, New York City.

Hardwick-Hindle, Inc., of 215 Emmet Street, Newark, N. J., have recently issued their new catalog No. 929 covering "Volume Faders" and catalog No. 429 on "Resistors." These may be had upon request.

A series of folders describing synthane, a laminated bakelite, may be obtained by addressing the Synthane Corporation, Oaks, Pa.

The Shallcross Mfg. Co. will be pleased to supply without charge copies of its catalog on Super Akra-Ohm, wirewound resistors which may be obtained in values up to five megohms. They may be addressed at 700 Parker Avenue, Collingdale, Pa.

The Supreme "Diagnometer" is described in a booklet issued by the Supreme Instruments Corporation of Greenwood, Miss.

"Instructions for Servicing Radio Receivers" is the name of a new booklet issued by the Jewell Electrical Instrument Co. of 1640 Walnut Street, Chicago, Ill. It may be obtained upon request and contains service data in chart form on more than a hundred standard commercial receivers.

Proceedings of the Institute of Radio Engineers
Volume 17, Number 12 *December 1929*

MONTHLY LIST OF REFERENCES TO CURRENT RADIO LITERATURE

THIS is a monthly list of references prepared by the Bureau of Standards and is intended to cover the more important papers of interest to professional radio engineers which have recently appeared in periodicals, books, etc. The number at the left of each reference classifies the reference by subject, in accordance with the scheme presented in "A Decimal Classification of Radio Subjects—An Extension of the Dewey System," Bureau of Standards Circular No. 138, a copy of which may be obtained for 10 cents from the Superintendent of Documents, Government Printing Office, Washington, D. C. The various periodicals can be secured from their publishers and can be consulted at large public libraries.

R100. RADIO PRINCIPLES

R113.4 Hulburt, E. O. Ionization in the atmosphere of Mars. PROC. I. R. E., **17**, pp. 1523–27; September, 1929.

(The composition of the atmosphere of Mars to great heights is calculated on certain assumptions and the electron density in the atmosphere due to the ultraviolet light of the sun is theoretically found. Calculated skip distances indicate 47 meters as the shortest wave for reliable long distance wireless communication over the surface of Mars. The improbability of interplanetary radio communication is pointed out.)

R113.4 Hafstad, L. R. and Tuve, M. A. Further studies of the Kennelly-Heaviside layer by the echo method. PROC. I. R. E., **17**, pp. 1513–23; Sept., 1929.

(Recent observations of the Kennelly-Heaviside layer by the echo method are reported. Two 24-hour series of observations showed a marked diurnal variation in the effective height of the layer and in the echo pattern received for each transmitted "peak." The pattern was most complex at night. Observations during magnetic disturbances showed an unusually great effective height and a change in the echo pattern.)

R113.5 Anderson, C. N. Notes on the effect of solar disturbances on transatlantic radio transmission. PROC. I. R. E., **17**, pp. 1528–35; Sept., 1929.

(The effects of magnetic storms on 60-kc transatlantic radio transmission and also on short-wave radio transmission are shown. In general individual storms tend to increase 60-kc daylight signal intensities and to decrease high-frequency signal field intensities.)

R113.6 Breit, G. Group velocity and long retardations of radio echoes. PROC. I. R. E., **17**, pp. 1508–12; Sept., 1929.

(van der Pol's hypothesis that group velocity may account for the retardations (up to 15 sec.) of echoes observed by Störmer is analyzed. It is shown that only under special circumstances can the electron distribution in the Kennelly-Heaviside layer be proper. A favorable condition is obtained if the refractive index decreases exponentially with the height.)

R113.6 Taylor, A. H. and Young, L. C. Studies of echo signals. PROC. I. R. E., **17**, pp. 1491–1507; September, 1929.

(Work reported in the PROC. I. R. E. for May, 1928, by the same authors has been continued as a more extended study of echo signals with particular reference to directional characteristics and to diurnal variations. Attention has been given to the relations between echo signals, frequency and the effective height of the Kennelly-Heaviside layer.)

R113.6 Pedersen, P. O. Wireless echoes of long delay. *Det. Kgl. Danske Videnskabernes Selskab* (Mathematical and Physics number, 48 pages), 9; 1929.

(Shows mathematically that radio echoes occurring after 10 seconds cannot be due to propagation of waves within the earth's atmosphere, that echoes occurring after intervals up to 30 seconds are due to propagation along or reflection from Störmer bands as explained in *Nature* (122, p. 681, 1928) that echoes after several minutes must be from outside the space in which the earth's magnetic field exerts appreciable effect. Transmissions at various wavelengths are also treated.)

R125 Zusammenfassender Bericht: Ueber Fehlweisung bei der Funkpeilung (Summary report: On errors observed in direction-finder bearings). *Zeits. fur Hochfrequenztechnik*, **34**, pp. 60–65; August, 1929.

R130 Chaffee, E. L. Equivalent circuits of an electron triode and the equivalent input and output admittance. Proc. I. R. E., **17**, pp. 1633–48; Sept., 1929.

(Two fundamental equivalence theorems concerning a triode and its circuits are derived. They give the single circuits which are equivalent to the plate and grid circuits of the triode, and contain only constant circuit elements and fictitious electromotive forces, the use of which greatly simplifies the calculation of currents in the triode circuits when the electrical variations are small. The equivalence theorems are used to obtain the equivalent input and internal output admittances of a triode with its associated circuits.)

R130.4 Rockwood, A. C. and Ferris, W. R. Microphonic improvement in vacuum tubes. Proc. I. R. E., **17**, pp. 1621–32; September, 1929.

(The causes and effects of microphonic disturbances in small receiving tubes is discussed. Methods of testing the microphonic qualities of tubes are described. A new type of tube with low microphonic output but with low filament power for use when microphonic troubles may be serious is also described and its characteristics are given.)

R131 Sowerby, A. L. M. The pentode as an anode rectifier. *Wireless Wld. and Radio Rev.*, **25**, pp. 283–286; Spet. 18, 1929.

(Curves, experimentally obtained are given to show the high sensitivity and large output obtainable through the use of the pentode as a plate rectifier. (To be concluded).)

R131 Colebrook, F. M. The triode valve equivalent network. *Experimental Wireless and W. Engr.* (London), **6**, pp. 468–87; September, 1929.

(From an analysis of the equivalent network of a tube expressions are developed for the input admittance and the voltage amplification in terms cf the known tube constants and the lead in the anode circuit. Deductions are made from the expressions for audio frequencies and for radio frequencies and are expressed in curves.)

R132 Möller, H. G. Berechnung des günstigsten Durchgriffes der Röhren in Widerstandsverstäcker. (Calculation of the optimum amplification factor for tubes used in resistance-coupled amplifiers.) *Zeits. fur Hochfrequenztechnik*, **34**, pp. 53–56; August, 1929.

(Theory and method of measurements.)

R133 Heegner, K. Ueber Schwingungserzeugung mittels eines Elektronenrohrensystems, bei welchem die Kapazitat von untergeordneter Bedeutung ist. (The production of oscillations by means of vacuum tube systems wherein the capacity is of subordinate importance.) *Zeits. fur Hochfrequenztechnik*, **34**, pp. 49–52; Aug., 1929.

(Description and theory of operation of two-tube circuit containing only *L* and *R*. Discussion of its peculiar properties is given.)

R134 Beers, G. L. Power detectors produce superior results in radio receivers. *Electric Journal*, **26**, pp. 413–415; September, 1929.

(Curves experimentally obtained are given to show the superiority of the negatively biased detector over the grid-leak-condenser detector in receivers using a power detector and a single stage of audio-frequency amplification.)

R140 Winter-Gunther, H. Ueber die selbsterregten Schwingungen in Kreisen mit Eisenkernspulen. (The self-induced oscillations in circuits containing iron-cored inductance.) *Zeits. fur Hochfrequenztechnik*, **34**, pp. 41–49; August, 1929.

(Description and mechanical analogy.)

R141.1 Pack, S. W. C. The frequency departure of thermionic oscillators from the LC value. *Experimental Wireless and Wireless Engr.* (London), **6**, pp. 472–80; September, 1929.

(The preliminaries of an investigation into the departure of the frequency of a tube a-c generator from the "LC" value under varying conditions of grid coupling, grid bias, filament current, plate voltage, and added resistance in the oscillatory circuit are described. The theory of the frequency of an electron tube generator is outlined.)

R150 Vincent, J. H. Experiments on magnetostrictive oscillators at radio frequencies. *Proc.* Phys. Soc. (London), **41**, pp. 476–86; August, 1929.

(An account is given of the behavior of two magnetostriction oscillators 6 mm and 4.5 mm in length, when placed in a coil in series with the main induction coil of a simple electron tube maintained oscillating circuit. The frequency characteristic of the smaller oscillator was 540 kc per sec.)

R162 Colebrook, F. M. The definition of selectivity. *Experimental Wireless and W. Engr.* (London), **6**, 422–24; August, 1929.

(The distinction between the selectivity of a radio circuit and its sharpness of tuning is discussed and a formulation is suggested. The application of the formulation to certain cases is made in an appendix.)

R170 Van Etta, E. and White, E. L. Radio interference from line insulators. *Jour.* A. I. E. E., **48**, pp. 682–86; September, 1929.

(Discussion of the causes of radio interference from insulators on high-voltage equipment. Present methods of eliminating this kind of disturbance are explained and the question of future design is discussed.)

R200. RADIO MEASUREMENTS AND STANDARDIZATION

R214 Harding, J. W. and White, F. W. G. On the modes of vibration of a quartz crystal. *Philosophical Mag.* (London), **8**, pp. 169–179; August, 1929.

(Four models showing the possible normal modes of vibration of a quartz crystal are constructed—a longitudinal and a transverse mode for each of the two directions of the alternating field. The patterns formed by lycopodium powder on the surfaces of the vibrating crystal are the bases of the construction.)

R300. RADIO APPARATUS AND EQUIPMENT

R344.3 Page, R. M. An investigation of the phenomena of frequency multiplication as used in tube transmitters. PROC. I. R. E., **17**, pp. 1649–55; September, 1929.

(An experimental investigation of the commonly used method for obtaining crystal control of high-frequency transmitters from crystals of lower frequency by means of a frequency multiplying vacuum-tube amplifier is described. Satisfactory operation of the amplifier is shown to depend upon the inductive reactance in the grid circuit which produces regeneration through the grid-plate feedback in the tube.)

R386 Hirsch, C. J. Notes on the design of r. f. band-pass filters. *Radio Engineering*, **9**, pp. 31–33; September, 1929.

(Mathematical treatment of band-pass filter design in reference to the screen-grid tube.)

R500. APPLICATIONS OF RADIO

R565 Hooper, S. C. Naval communications—Radio Washington. PROC. I. R. E., **17**, pp. 1595–1620; September, 1929.

(The radio facilities of Radio Washington are described. There is included a brief description of receiving equipment, and of methods of control of transmitters in use at the Navy Department, Radio Central, and of the transmitting equipment installed at Arlington, Va., and Annapolis, Md.)

R582 Jenkins, C. F. The drum scanner in radiomovies receivers. PROC. I. R. E., **17**, pp. 1576–1583; September, 1929.

(A description is given of the Jenkins drum scanner for use in television. Its advantages over the disk scanner are pointed out.)

R582 Horton, J. W. The electrical transmission of pictures and images. PROC. I. R. E., **17**, pp. 1540–63; September, 1929.

(Some of the important principles underlying the electrical transmission of pictures are discussed. The subject is treated as a special form of electrical communication. The dependence of the rate at which information is transmitted upon the frequency range occupied by the signal is shown to be of fundamental importance. The relations between the original picture and the transmitted signal as it appears in the several factors of the system are studied.)

R582 Goldsmith, A. N. Image transmission by radio waves. PROC. I. R. E., **17**, pp. 1536–39; September, 1929.

(A general outline of the problems of facsimile transmission and television, and their development, introductory to papers presented at the 1929 Institute of Radio Engineers Convention in Washington, D. C.)

R582 Ranger, R. H. Mechanical developments of facsimile equipment. PROC. I. R. E., **17**, pp. 1564–75; September, 1929.

(Recent mechanical developments in equipment for photoradio facsimile transmission are described.)

R582 Weinberger, J., Smith, T. A., and Rodwin, G. The selection of standards for commercial radio television. PROC. I. R. E., **17**, pp. 1584–94; September, 1929.

(The basis of a system of television standards suitable for commercial television service with specific standards is discussed. The elements considered in this paper are: picture proportion, number of scanning elements, number of picture repetitions per second, scanning method and direction, and phase of transmitted current.)

R600. RADIO STATIONS: EQUIPMENT, OPERATION AND MANAGEMENT

R613 Lamb, J. J. WTIC, a modern 50-kw broadcast station. *QST*, **13**, pp. 9–16; October, 1929.

(Description of a modern broadcast station.)

R800. NON-RADIO SUBJECTS

R537.1 Biedermann, E. A. Potential difference and electromotive force. *Experimental Wireless and W. Engr.* (London), **6**, pp. 481–485; September, 1929.

(A discussion of the definition of "potential difference" proposed by R. M. Wilmotte (*Experimental Wireless and W. Engr.*, November, 1928) is presented. Faults are found and it is proposed to generalize the definition of potential difference between points as the line integral between those points of only that force which depends on the charge distribution.)

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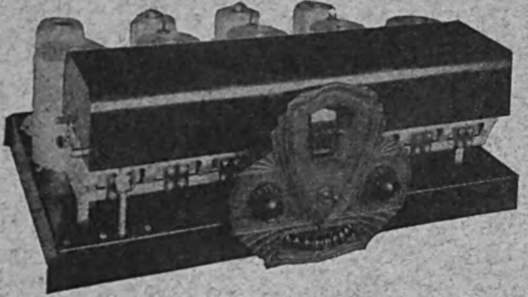

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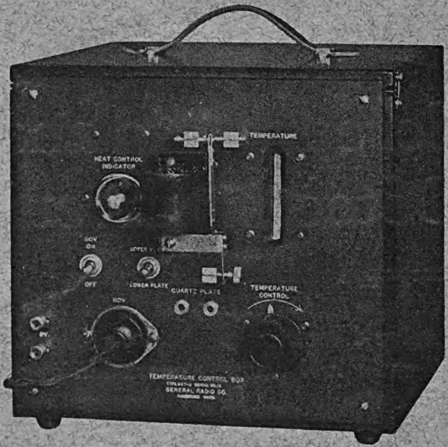
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